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1919

EDITED BY

ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY

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THE INSTITUTE OF RADIO ENGINEERS
announces with regret the death of Messrs.

Edward J. Baskin
Guy E. Morse
William D. Woodcock

Edward J. Baskin was born in Lynn, Massachusetts in 1899. He was a graduate of the Chelsea Grammar School, the Northeastern Preparatory School, the Eastern Radio School, and the Gillespie Aviation School of Boston. During 1917 he was employed as operator on coastwise ships plying to the West Indies by the Marconi Wireless Telegraph Company of America. In 1918 he enlisted in the Naval Reserve, Aviation Section. After being stationed at Charleston, South Carolina, he contracted pneumonia, and died on October 8, 1918. He was a member of The Institute of Radio Engineers.

Guy E. Morse was the son of Mr. and Mrs. Ernest R. Morse of Kansas City, Missouri. He was born in 1895 at Wolfville, Nova Scotia, his parents being then Canadians. He was a graduate of the Kansas City public schools, and had completed two years of his studies at the University of Illinois where he was preparing himself for electrical engineering. He enlisted in the first Officers' Training Camp, at Fort Sheridan, Illinois in 1917, and was transferred to Fortress Monroe. He was there commissioned as Second Lieutenant in the Coast Artillery, and stationed at Key West. On application, he was transferred to the air service as an aerial observer, and took his ground training at Austin, Texas. In 1918, he was sent to France, and was trained there at Saumur, Tours, and Cazaux, being then attached to the 135th Aero Squadron. In August, 1918 he was sent to the front, and was killed in aerial combat at St. Mihiel, September 12, 1918. He died at the age of twenty three, having been cited for gallantry before his death. He was well known as an excellent student and a fearless young man.

William D. Woodcock was born in 1896. He was graduated from the Lafayette High School, and was a member of the class of 1919 in analytical chemistry at the University of Buffalo. Before the war, he was a well known radio amateur, a member of an organization of radio operators, a member of The Institute of Radio Engineers, holder of a first class communication license, and of a special station license. He was the Buffalo operator who transmitted the President's message sent from coast to coast on October 27, 1916. After war was declared, he enlisted in the Naval Reserve, was sent to Great Lakes in 1917, and appointed a first-class operator, first at Cleveland, and then at Buffalo. After several service changes, he was made an instructor in the Radio School at Great Lakes, and then advanced to the radio laboratory. There he contracted pneumonia, which caused his death.

RESONANCE MEASUREMENTS IN RADIOTELEGRAPHY WITH THE OSCILLATING AUDION*

By

LOUIS W. AUSTIN

(UNITED STATES NAVAL RADIOTELEGRAPHIC LABORATORY, WASHINGTON, D. C.)

For purposes of rough tuning, many workers have doubtless made use of the click heard in the telephones of an oscillating audion circuit when it is brought into resonance with another circuit at proper coupling. As the resonance click has apparently not been mentioned in any of the publications on radio frequency measurements, it seems probable that it is not generally known that this click offers by far the quickest and simplest means of making nearly all measurements depending on the determination of resonance. The accuracy is quite equal to that obtainable with sensitive thermo-elements, and greatly superior to the accuracy of the detector and telephone method.[1]

Since the audion circuit itself is not suited to exact calibration, the substitution method is generally used. The following examples illustrate the procedure:

CAPACITY OF AN ANTENNA BY SUBSTITUTION

The antenna is loaded with inductance so as to give a wave length of five to ten times the fundamental, then the oscillating audion circuit is coupled to the antenna inductance, and the audion tuning condenser varied until a click is heard in the telephones. In general, if the coupling is close, the click will be heard at different points with increasing and decreasing condenser capacity. The coupling should then be loosened until both clicks appear at the same condenser setting, or, if this is impossible, the mean setting is taken provided the points are less than a degree apart. Next, leaving the audion condenser on the resonance point, the ground and antenna are discon-

* Received by the Editor, July 24, 1918.

[1] Care must be taken regarding harmonics in all measurements in which bulbs are used for excitation.

nected from the antenna inductance and replaced by the calibrated variable substitution condenser. This last is adjusted to resonance with the audion circuit exactly as described above, and the capacity of the condenser is then equal to that of the antenna, subject to a small correction for the natural antenna inductance.

WAVE LENGTH OF A DISTANT STATION

The receiving antenna and secondary oscillating circuit are first tuned exactly to the distant station, preferably at loose coupling, the audion tuning condenser being adjusted to give the dead point of the beats in the case of continuous wave reception. Next, without changing anything in the antenna or secondary, a wave meter is coupled to the secondary and adjusted to resonance by the click method. The reading of the wave meter gives at once the wave length for the sending station.

In a similar way, wave meters can be compared and condensers and inductances calibrated, either by substitution or by making use of the well-known relation existing between the product of inductance and capacity and the wave length.

Besides the simplicity and quickness of this method, it has the advantage that it does away with the necessity for all auxiliary apparatus in the wave meter, and enables measurements of the highest accuracy to be taken on shipboard and in other places where the use of sensitive galvanometers is impossible.

U. S. Naval Radiotelegraphic Laboratory.
June, 1918

SUMMARY: The telephone click in an oscillating audion circuit when a coupled circuit is brought into tune with it is utilized to measure quickly and accurately antenna capacity, wave length of distant stations, capacities, inductances, and wave lengths.

A BRIEF TECHNICAL DESCRIPTION
OF THE
NEW SAN DIEGO, PEARL HARBOR, AND CAVITE HIGH POWER NAVAL RADIO STATIONS*

(Supplementing Captain Bullard's Paper)

By

Leonard F. Fuller

(Chief Electrical Engineer, Federal Telegraph Company)

As mentioned in Captain Bullard's article, 11,000-volt, 3-phase, 60-cycle power will be delivered to the new San Diego high power radio station. The power equipment is being manufactured by the General Electric Company and will consist of the usual oil switches, switchboards, and motor-generators used in power work.

The motor-generators, which are in duplicate, will be 2-unit, 4-bearing sets, consisting of 300 horse-power, 1,200 revolutions per minute, 2,200-volt, 3-phase, 60-cycle, squirrel cage induction motors, direct connected to 200 kilowatt, 1,200 revolutions per minute, 950-volt, compound wound, direct current generators with 2-kilowatt, 125-volt, over-hung exciters.

The temperature rises of the motors will be 40° C. for continuous full load operation and not over 55° C. for a continuous series of duty cycles of 1.5 hours on and 1 hour off at 125 per cent. of load.

The generator temperature rises will be 40° C. for continuous full load operation and not over 55° C. for a continuous series of duty cycles of 1.5 hours on and 1 hour off at 250 kilowatt output. The sets as a whole can withstand a 100 per cent. overload for short periods.

Duplicate 14-kilowatt, 125-volt, direct current, motor-generators with a complete set of spare parts will be installed also, for furnishing power for plant auxiliaries.

The Federal-Poulsen arc converter will be of the oil-immersed, water-cooled type capable of furnishing 150 amperes radiation continuously, and 170 amperes for periods of 1.5 hours on and 1.5 hours off. The temperature rises will be 40° C. and 50° C. respectively for all current-carrying or electrical parts.

* Received by the Editor, May 20, 1916.

At Pearl Harbor, power will be supplied, as Captain Bullard states, at 2,200 volts, 3-phase, 60 cycles. This plant is similar to that at San Diego but of higher power. The large motor-generators are provided in duplicate with a complete set of spare parts including spare armatures. They are 2-unit, 4-bearing sets, manufactured by the General Electric Company, consisting of 750 horse-power, 900 revolutions per minute, 2,200-volt, 3-phase, 60-cycle, wound rotor, induction motors, direct connected to 500 kilowatt, 900 revolutions per minute, 1,430-volt, compound wound, direct current generators with 3-kilowatt, 125-volt, compound wound, over-hung exciters.

The temperature rises of both motors and generators will be 50° C. on a continuous series of duty cycles of 2 hours on and 1 hour off. High voltage direct current switchboards and control apparatus of a type used in electric railway work will be provided.

The Federal-Poulsen arc converter will be of the oil-immersed, water-cooled type, capable of furnishing 200 amperes radiation continuously and 250 amperes for periods of 1.5 hours on and 1.5 hours off with temperature rises of 40° C. and 50° C. as specified for San Diego.

On account of the oil immersion and water cooling of all coils and arc converter windings, this type of apparatus is extremely rugged and reliable. The arc transforms or converts the 1,500-volt direct current power into radio frequency energy without the use of rotating parts or radio frequency magnetic circuits.

The antenna loading coil will be of litzendraht cable supported entirely by porcelain. This cable will be 1.75 inches (4.45 cm.) in diameter, and the loading coil diameter will be 14 feet (4.26 m.).

The wave changer will be of the rotary type capable of throwing the plant onto any one of five wave lengths when operating at full power. This feature holds for San Diego and Cavite as well as Pearl Harbor and in all three the usual antenna grounding switches, transfer switches, etc., will be provided.

The generating equipment at Cavite will be identical with that at Pearl Harbor except that the power supply will be 220-volt direct current and no over-hung exciter will be used. The 500-kilowatt, 1,500-volt, generator excitation will be supplied from the 220-volt direct current buses. At both Pearl Harbor and Cavite, special insulation and fittings for tropical conditions will be provided.

The arc and the remainder of the radio set at Cavite will be in every way identical with that at Pearl Harbor.

The arcs at Pearl Harbor and Cavite will be approximately 9 feet 2.5 inches (2.81 m.) by 7 feet 4 inches (2.23 m.) wide by 12 feet (3.66 m.) long, and will each weigh approximately 60 tons (54,000 kg.) under operating conditions. The San Diego arc is somewhat smaller and will weigh 21 tons (19,000 kg.) under operating conditions. The larger arcs will present an outside appearance very similar to certain types of vertical shaft hydro-electric generators.

San Francisco, California, May 12, 1916.

SUMMARY: The new high power Federal-Poulsen arc stations of the United States Navy are described briefly. The motor-generator sets, temperature rises, field excitation, and operating characteristics are given. The arc-converter, antenna loading inductance, and wave changing switch are described shortly.

(The distance from Cavite, Philippine Islands to Pearl Harbor, Hawaii, is 5,300 miles (8,500 km.), from Pearl Harbor to San Diego 2,600 miles (4,200 km.), and from Cavite to San Diego 7,800 miles (12,500 km.), altogether over water. The approximate power used to cover 4,000 kilometers is 70 kilowatts in the antenna, and for 8,500 kilometers is 160 kilowatts. The former value may be compared with that given by Mr. John L. Hogan, Jr., on page 419 of the October, 1916, issue of the PROCEEDINGS, namely 72 kilowatts.—EDITOR).

HYSTERESIS AND EDDY-CURRENT LOSSES IN IRON AT RADIO FREQUENCIES*

By

Christian Nusbaum

(Jefferson Physical Laboratory, Harvard University, Cambridge, Massachusetts)

Introduction

The study of the heat losses due to hysteresis and eddy-currents in iron, when subjected to an oscillating magnetic field, is of practical interest in radio telegraphy and radio telephony. In this paper will be described a calorimetric method for the determination of these losses, in a certain specimen of soft iron, at frequencies between the limits of 150,000 and 1,000,000 cycles per second, and the results obtained by this method of measurement will also be stated.

Various methods of measurement were used by previous investigators for the determination of these losses, and the results obtained are quite at variance with each other. The results of Warburg and Honig,[1] Tanakadate,[2] and Weihe[3] indicate that the loss per cycle, at frequencies of three cycles and sixty cycles per second, is less than for static magnetization. The results of Hopkinson,[4] Maurain,[5] Gray,[6] and Guye and Herzfeld[7] indicate that the loss per cycle is independent of the frequency, while those of Steinmetz,[8] Niethammer,[9] M. Wien,[10] Krogh and Rikla,[11] Dina,[12] and Corbino,[13] lead to the conclusion that the loss per cycle increases with the frequency.

*Received by the Editor, July 10, 1918.

[1] Warburg and Honig, Wied. Ann., 20, p. 814, 1883.
[2] Tanakadate, A., Phil. Mag., 28, p. 207, 1889.
[3] Weihe, F. A., Wied. Ann., 61, p. 578, 1897.
[4] Hopkinson, Elektrotechn. Zeitschr., 13, p. 642.
[5] Maurain, C., Ecl. Electr., 15, p. 499, 1898.
[6] Gray, Phil. Trans. 174 A, p. 351.
[7] Guye and Herzfeld, Compt. Rend., 136, p. 957, 1903.
[8] Steinmetz, Elektrotechn. Zeitschr., 13, p. 957.
[9] Niethammer, Wied. Ann., 66, p. 29, 1898.
[10] Wien, M., Wied. Ann., 66, p. 859, 1898.
[11] Rikla and Krogh, Elektrotechn. Zeitschr., p. 1083, 1900.
[12] Dina, A., Elektrotechn. Zeitschr., p. 41, 1902.
[13] Corbino, O. M., Atti. Assoc. Electr., Ital., 1903.

Alexanderson,[14] using a radio frequency generator, obtains results which may be stated as follows: (1) that the permeability is the same at high as at low frequencies, and (2) that a higher induction than 1,200 lines per square centimeter cannot be reached on account of the "shielding effect" due to eddy-currents. The loss per cycle is less for higher than for lower frequencies, the observations extending from 40,000 to 200,000 cycles per second. Recently, Fassbender and Hupka[15] outlined an experimental method for the determination of the permeability and the form of the hysteresis cycle for very high frequencies. Preliminary observations clearly show the change in the form of the hysteresis cycle due to the "shielding effect" of eddy-currents. The results also show, that even for a frequency of 200,000 cycles per second, it is impossible to obtain wire of sufficiently small diameter to prevent disturbing effects due to eddy-currents.

Method and Apparatus

If an undamped oscillatory current is sent thru the winding of a toroid containing a core of soft iron wires, the iron core will, during each oscillation, pass thru a complete hysteresis loop. On the other hand, if the current is damped, however little, there will be a series of non-closed hysteresis loops for each train of oscillations. The rapid change in the magnetic induction will induce eddy-currents, which in turn cause a non-uniform distribution of the magnetic flux thru the cross-section of the iron. These processes manifest themselves in heat, and are generally designated as "heat losses."

For the purpose of the measurement of these losses, two toroids, both being alike in every respect except that only one contains a specimen of the iron the losses of which are to be determined, are connected in series with each other in the secondary of an oscillating circuit. In Figure 1, T_o and T, respectively, represent the toroids without and with iron. Each of these toroids is placed in a separate calorimeter. If the heat capacity of the calorimeters, as well as of the toroids, is the same, and the $I^2 R$ losses in the winding of the two toroids are equal, then the heat produced by the iron specimen in one calorimeter may be balanced in the other one by means of a direct-current heating coil ($H C$, Figure 1).

Each calorimeter consists essentially of two parts, an inside vessel C_i, Figure 2, and an outside one, C_o. The inside vessel

[14] Alexanderson, Elektrotechn. Zeitschr., 32, p. 1078, 1911.
[15] Fassbender and Hupka, Jahrbuch der Drahtlosen, Telephonie und Telegraphie., 6, p. 133, 1913.

is supported by the brass cones *c* and the ebonite disc *r*, and separated from the outside vessel by an air space and the three equally-spaced insulators *i*, only one of which is shown in the drawing. Both vessels have their entire surfaces nickel-plated and polished. By such an arrangement the loss of heat by conduction and radiation is reduced to a minimum. The cover of each calorimeter is fitted to the outside vessel by a ground joint, and is provided with six vertical tubes. The central tube, 1, admits

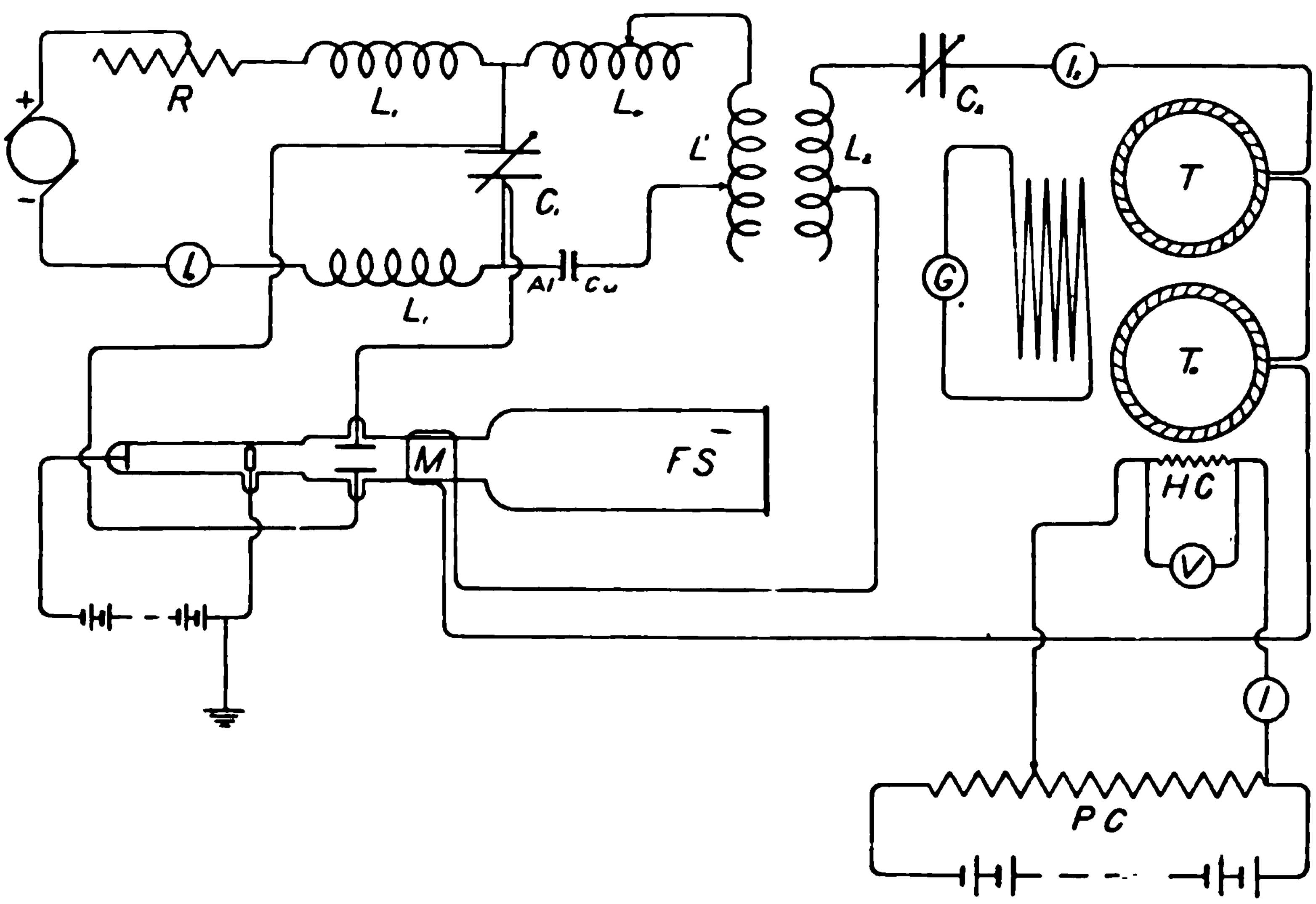

FIGURE 1—Diagram of Connections

one end of the thermo-pile; tubes, 2, admit the supports for the stirrers, *s*, and tubes, 3, the leads to the toroids. The remaining tube, 4, admits the leads of the direct-current heating coil which is placed in the calorimeter containing the toroid without iron. Each lead wire is carefully insulated from the tube which admits it. The stirrer in each calorimeter consists of two parallel discs, equal in diameter and held together by four uprights. Each disc has a series of circular openings arranged in the form of a circle. Its internal diameter is large enough easily to permit motion of the stirrers past the toroid, which is supported in the central portion of the calorimeter. Both of the calorimeters are mounted on a common base and entirely immersed in a bath of kerosene, so that the liquid surface is above the ground joints

but below the upper ends of the vertical tubes. The temperature of this external bath is maintained approximately the same as that of the calorimetric liquid within the calorimeters by means of a heating coil and stirring apparatus. The turns of the toroids are wound on two glass cylinders, about four centimeters (1.6 inch) in diameter and one centimeter (0.4 inch) in width, both cylinders having been cut from the same glass tubing.

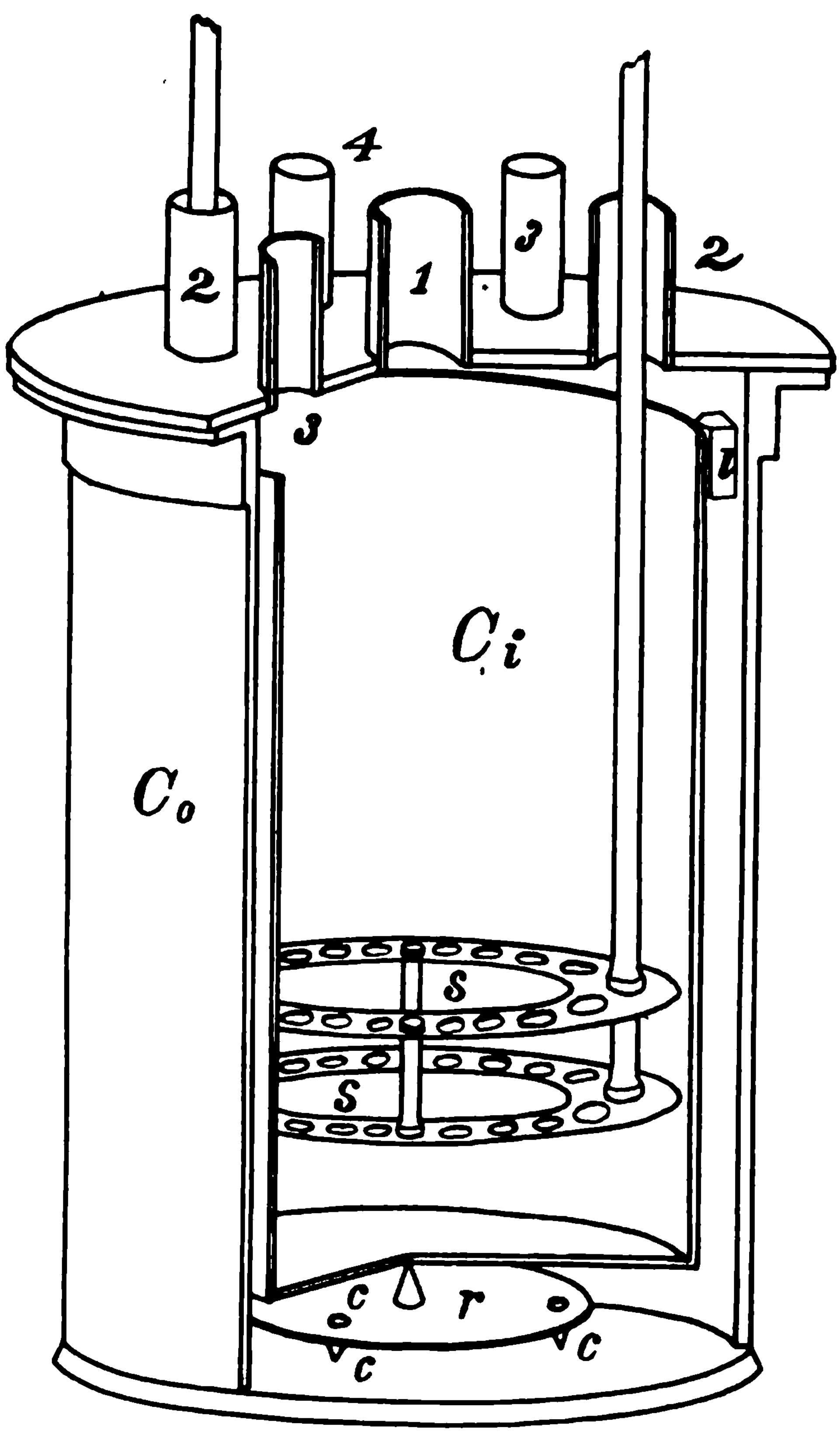

FIGURE 2—Diagram of Calorimeter

Around one of these cylinders the specimen of soft iron wire had been previously wound in the form of a helix of small pitch. With this type of core the individual turns of the specimen are well insulated from each other and their direction is nearly parallel to the direction of the magnetic field. A small quantity of iron is used in order that the inductance due to the presence

of the iron shall be a small fraction of the total inductance of the secondary circuit, and that there may not be an excessive rise in temperature within the calorimeter.

The heating coil, made from number 30 constantan wire,* is wound non-inductively on an ebonite support. The current thru the coil can be varied by "taking off potentials" from the potentiometer $P\,C$, Figure 1. The ammeter I is in the circuit of the heating coil, and the voltmeter is connected across its terminals.

To test for the equality of temperature between the two calorimeters, a thermo-pile consisting of 12 copper-constantan junctions is used. These junctions are arranged in two groups of six junctions each. With such an arrangement the two groups may be balanced against each other and thus the circuit tested for stray electromotive forces. To reduce the latter to a minimum all terminal contacts are of copper, as well as all switches, and the lead wires are well sheathed and protected.

The Chaffee [16, 17] gap, used as the source of the oscillations, possesses the especial advantage of not only producing continuous oscillations of a definite wave form at radio frequencies, but also of operating in such a manner that the oscillogram of its current wave can readily be obtained. On account of its method of operation, the damping of the current is generally very small. It consists, essentially, of an aluminum cathode and an anode of copper, both terminals being surrounded by an atmosphere of hydrogen. The general diagram of the circuit is shown in Figure 1. The primary circuit consists of an e m f. of 500 volts, direct-current, connected in series with a variable resistance R, two choke coils L_1, the gap G, and two variable inductances L' and L_o. In parallel with the inductances L' and L_o and the gap itself is placed a variable condenser C_1. The secondary circuit consists of the variable inductance L_2, air condenser C_2, hot-wire ammeter I_2, the two toroids T and T_o, and the magnetic deflecting coil M. The variable inductance L' of the primary and the variable inductance L_2 of the secondary circuit form a closely connected oscillation transformer. The linkages of the primary thru the secondary may be changed by varying the number of turns of the primary inductance. The Braun tube is used in this experiment to indicate the condition of complete syntony between the primary and

*Diameter of number 30 wire = 0.0100 inch = 0.025 cm.

[16] Chaffee, E. L., Proc. Am. Acad. Arts and Sci., 47, p. 267, 1911.

[17] Washington, Bowden, PROC. INST. RADIO ENGRS., 4, p. 341, 1916.

the secondary circuits, and also to indicate the form of the current wave in the secondary circuit. The terminals of the primary capacity C_1 are connected to the electrostatic plates of the tube. The magnetic deflecting coil M, Figure 1, is a part of the secondary circuit, and is used to deflect the cathode beam at right angles to the deflection of the electrostatic field.[18] This coil consists of sixteen turns of annunciator wire (number 18, cotton covered*), half of the turns being on opposite sides of the tube. Electrostatic deflection of the cathode beam by the coil is avoided by completely surrounding the tube inside the coil by a split solenoid.[19] The fluorescent screen $F\,S$, Figure 1, made from finely powdered calcium tungstate, was used for visual observation and for photographic purposes, and was found very satisfactory. The camera used in photographing the oscillogram was placed below the screen, so that its axis coincided with the axis of the tube. By this arrangement distortion of the picture of the oscillogram is prevented.

Method of Operation and Procedure

The operation of the gap can be most readily understood by tracing the sequence of phenomena from the instant at which the system is started. When the potential of the primary condenser C_1 has attained a value sufficient to break down the high resistance of the gap, the discharge of C_1 and the main current I_o rush across the gap and thru the primary inductance L'. After the discharge, the main current remains constant and flows into C_1 at a practically constant rate, neutralizing the inverse charge on the condenser and charging it again in the initial direction. The secondary thus receives periodic impulses from the primary. The frequency of these impulses depends on the main current and the capacity in the primary circuit, and not on the duration of the discharge. During the intervals between the successive discharges of the primary circuit, the secondary circuit oscillates with its own free period and with a damping determined wholly by the conditions existing in it. The amplitude of the oscillations can be maintained nearly constant by making the primary impulses occur at every three or four oscillations of the secondary. Since the primary condenser charges up at a uniform rate, if its terminals are connected to the electro-

*Diameter of number 18 wire $=0.0403$ inch $=0.102$ cm.

[18] As drawn, the coil is in the plane of the paper and the deflection caused by it would be parallel to the deflection produced by the electrostatic field.

[19] Chaffee, E. L., above reference.

static plates of the Braun tube, and the oscillatory current in the secondary is run thru the magnetic deflecting coil, an oscillogram will be obtained on the fluorescent screen. This oscillogram indicates the wave form of the secondary current, and also shows the number of its oscillations for each discharge of the primary.

Preparatory to the taking of a set of observations, the capacity of the secondary circuit is varied until its frequency of oscillation has the value at which one wishes to make an observation. If the two circuits are not in syntony, the constants in the primary are varied until such condition is attained. The primary capacity is generally kept as large as possible, consistent with the energy to be transmitted and the number of oscillations of the secondary for each discharge of the primary. The current in the secondary is varied by changing the linkages between the inductances L' and L_2. During a set of observations for a given frequency, generally, the inductance L_2 is not altered, but the number of linkages changed only by varying the number of turns of the inductance L'. Since L' is used as one side of a variable transformer it is convenient to have an additional variable inductance L_0 in the primary.

The number of oscillations of the secondary per second is determined by means of a wave meter. The resonance between the secondary circuit and the wave meter was in most of the observations very sharp.

In the taking of any given observation, the current thru the direct-current heating coil is varied until the rate of rise of temperature in the two calorimeters is the same, as indicated by zero deflections of the galvanometer in the thermo-pile circuit, if the calorimeters are at the same temperature, or by a constant deflection of the galvanometer if the calorimeters are at a slightly different temperature. Generally, two preliminary observations are taken, one which will give a more rapid rise of temperature in the calorimeter with the heating coil, and a second which will give a less rapid rise of temperature. By interpolating between these two values of the current, a new and more accurate value can now be obtained. The calorimeters are then brought to the same temperature, the current set at the interpolated value, and the observation again repeated. If it is satisfactory, the heat developed by the heating coil in one calorimeter, and by the specimen of iron in the other calorimeter, can be calculated from the readings of the ammeter and voltmeter in the heating coil circuit.

Data, Results, Errors and Conclusions

The galvanometer, G_o, Figure 1, used in the thermo-pile circuit was a high resistance D'Arsonval instrument. Tho it was not especially adapted for thermo-pile work, and for the particular junctions used, yet a difference of temperature of 0.004°C. between the two calorimeters could be read directly, when the scale was at a distance of two meters (78.7 inches) from the galvanometer. During any given set of observations, there was in the majority of cases a rise of from 4 to 6 degrees within each calorimeter.

The thermal-junctions were arranged in two sets of six pairs each. By such an arrangement, even tho the calorimetric fluids were at different temperatures, there should be no deflection of the galvanometer if the two sets were opposed to each other, provided there is no other source of emf. in the circuit. At the beginning of every set of observations, the two sets of thermal-junctions were balanced against each other, as well as at the end of each set of observations.

The hot-wire ammeter I_2, in the secondary circuit, was

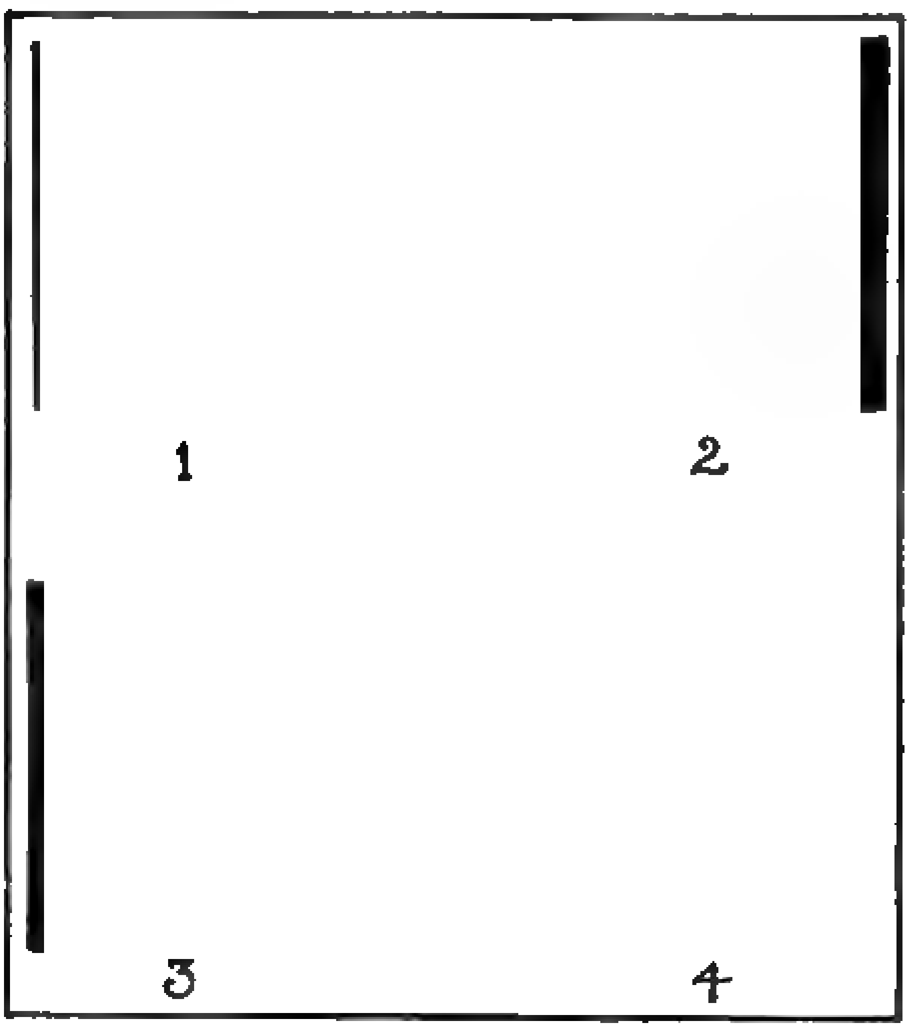

Plate 1

calibrated by comparison with direct-current values. In the calibration it was assumed that the resistance of the hot-wire ammeter does not vary appreciably with the frequency, so that the calibration curve will hold equally well for radio and for audio frequencies. This condition holds only when the wire of the ammeter is not more than 0.35 mm. (0.014 inch) in diameter. To serve as a check, comparisons were made on the Braun tube between the deflections produced by a direct-current of a given value and that produced by a radio frequency current. Figure 4 of Plate 1 shows such a comparison. The central spot indicates zero deflection.

The volume of the specimen of iron on the toroid T is 0.0298 cc. (0.0019 cu. inch). Both toroids are 4.078 cm. (1.61 inch) in diameter and the current-bearing winding of each consists of

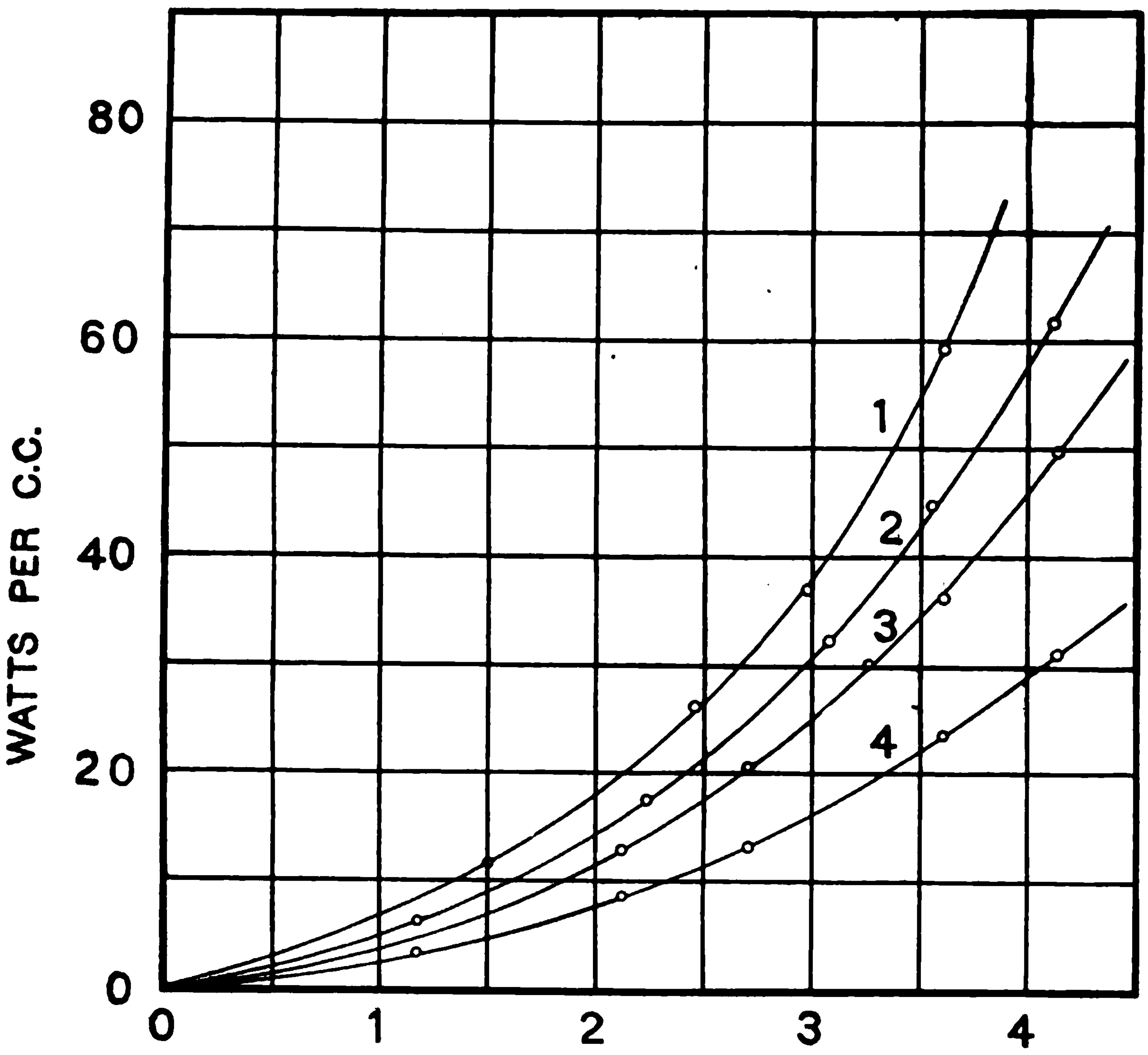

FIGURE 3—Curves Showing Heat Developed in Watts Per Cc. Plotted Against the "Effective Ampere-turns Per Centimeter"

Curve 1, 1,000,000 cycles
2, 550,000 "
3, 300,000 "
4, 150,000 "

72.5 turns. The term "effective ampere-turns per centimeter" is used to designate the product NI, where N is the number of turns per centimeter in the toroids and I the "square-root mean-square" values of the current as indicated by the hot-wire ammeter.

Figure 3 represents the results of Table I. The "effective ampere-turns per cm." as abscissas are plotted against the heat developed in watts per cc. as ordinates for the various frequencies.

TABLE I

Current Values

Deflection of hot-wire ammeter	Square-root mean-square values	Effective ampere-turns per cm.	Energy developed in watts	Energy developed in watts per cc.	Frequency
2.20	0.266	1.505	0.359	11.97	1,000,000
4.00	.433	2.451	.790	26.33	1,000,000
5.00	.526	2.977	1.112	37.07	1,000,000
6.20	.638	3.611	1.776	59.20	1,000,000
1.60	.207	1.172	0.184	6.13	550,000
3.60	.395	2.236	.528	17.60	550,000
5.20	.545	3.085	.974	32.47	550,000
6.10	.629	3.560	1.349	44.97	550,000
7.10	.724	4.097	1.850	61.67	550,000
3.40	.376	2.128	0.390	13.00	300,000
4.50	.479	2.711	9.626	20.87	300,000
5.60	.582	3.294	0.908	30.27	300,000
6.20	.638	3.617	1.089	36.30	300,000
7.20	.733	4.148	1.496	49.87	300,000
1.60	.207	1.172	0.101	3.35	150,000
3.40	.376	2.128	0.259	8.65	150,000
4.50	.479	2.711	0.404	13.45	150,000
6.20	.638	3.611	0.709	23.65	150,000
7.20	.733	4.148	0.937	31.25	150,000

In Table II are given the heats developed in ergs per cycle per cc., for definite values of the "effective ampere-turns per cm." These values are represented by the graphs of Figure 4.

Plate I shows the current-wave forms for various frequencies. The current in the secondary is thus not undamped, but the value of the damping can be made very small by making the primary impulses take place at every four or five oscillations of the secondary. This quantity is called by Chaffee [20] the "in-

[20] Chaffee, E. L., previous reference.

TABLE II

Effective ampere-turns per cm.	Energy developed in watts per cc.	Energy developed in ergs per cycle per cc.	Frequency
1	2.13	1.42×10^2	150,000
2	7.42	4.94×10^2	150,000
3	16.46	10.97×10^2	150,000
4	29.45	19.60×10^2	150,000
1	3.73	1.24×10^2	300,000
2	11.64	3.88×10^2	300,000
3	24.92	8.31×10^2	300,000
4	45.70	15.23×10^2	300,000
1	5.13	0.93×10^2	550,000
2	14.25	2.59×10^2	550,000
3	30.42	5.53×10^2	550,000
4	58.62	10.66×10^2	550,000
1	7.13	0.71×10^2	1,000,000
2	18.50	1.85×10^2	1,000,000
3	37.90	3.79×10^2	1,000,000
4	81.20	8.12×10^2	1,000,000

verse charge frequency." The value of this constant was 5 thruout the entire experiment. Table III gives the values of the frequency for the different figures of Plate I.

TABLE III

Figure..............	1	2	3	4
Frequency..........	150,000	550,000	300,000	1,000,000

Tests were made for the relative heat capacities of the calorimeters, differences of heat losses of the calorimeters due to radiation and losses due to heat conduction along the leads, and relative $I^2 R$ losses in the windings of each toroid.

These tests were made by sending a direct-current of constant value thru the windings of the two toroids and then observing the difference of the rate of increase in temperature between the two calorimeters. These tests indicated that the errors thus introduced were in the majority of the observations less than one-tenth of one per cent. The largest errors entering into the observations are the variations of the current values in the secondary circuit. During the time of an entire set of observations the current may vary from three to five per cent. of its mean value.

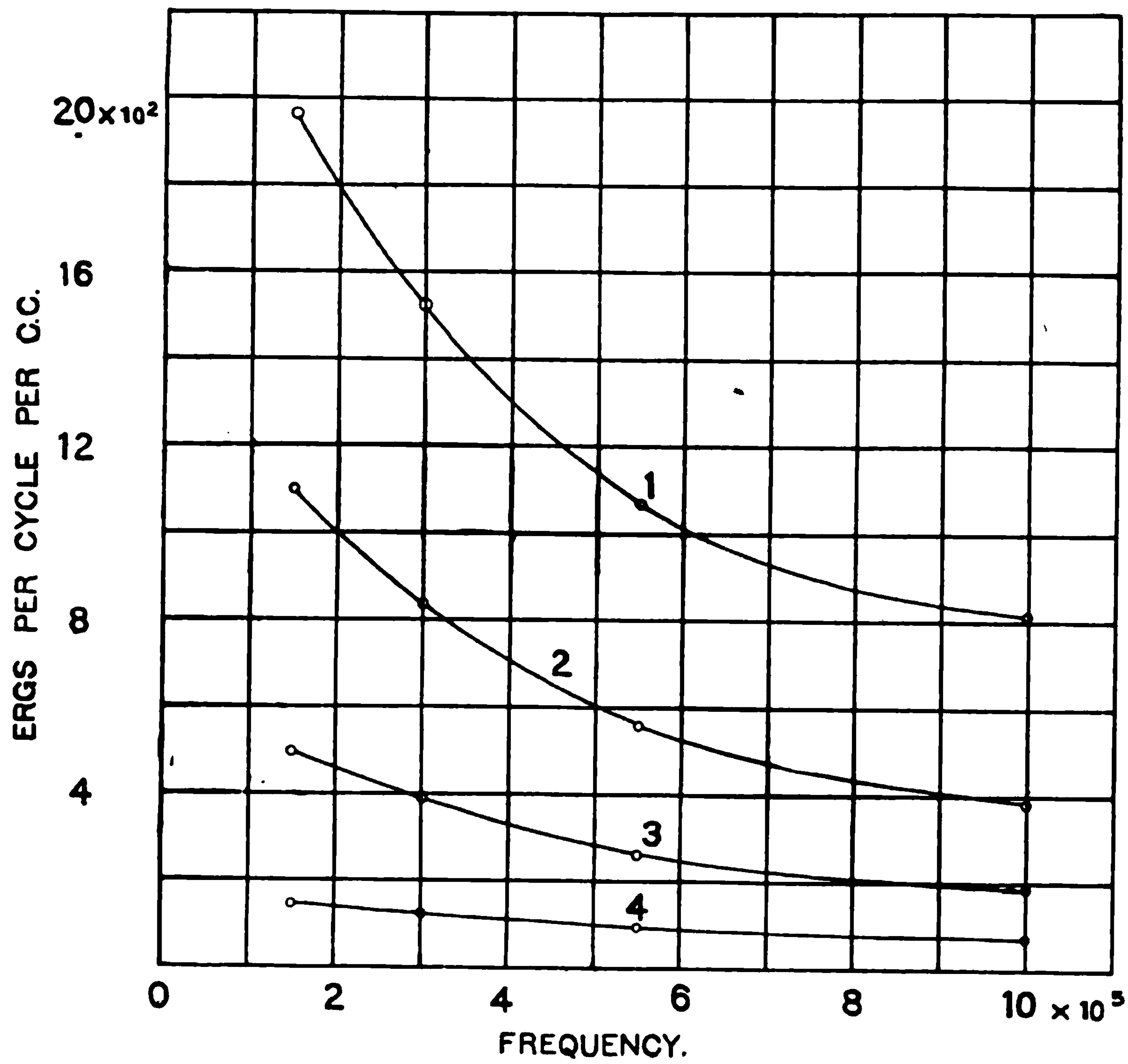

FIGURE 4—Curves Showing the Heat Developed in Ergs Per Cycle Per CC. Plotted Against the Frequency

Curve 1, 4 "effective ampere-turns per cm."
2, 3 " " " " "
3, 2 " " " " "
4, 1 " " " " "

The results of the experiment show that the heat developed per cycle by magnetic hysteresis and eddy-currents in iron decreases with an increase in the frequency of the oscillating magnetic field. This fact indicates a decrease of the magnetic induction on account of the "shielding-effect" produced by the eddy-currents.

Jefferson Physical Laboratory,
Harvard University,
June 1, 1918.

SUMMARY: After reviewing the bibliography of heat losses per cycle at various frequencies, the author describes a comparison calorimetic method whereby the losses in the soft iron wire core of a toroid are measured (against a similarly wound toroid without an iron core). The Chaffee gap radio frequency generator is used. Braun tube oscillograms of the cycle are shown, together with experimental data on the iron losses. It is found that the loss per cycle decreases as the frequency increases.

THE MEASUREMENT OF RADIO FREQUENCY RESISTANCE, PHASE DIFFERENCE, AND DECREMENT*

By
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Various methods of measuring radio (high) frequency resistances, the power factor or phase difference of radio condensers, the sharpness of a resonance curve, and the decrement of a wave have been given. It is proposed in this paper: (1) to show that these quantities all express essentially the same physical magnitude and hence a single process of measurement gives them all; (2) to derive and classify the methods of measurement; (3) to describe improvements in these measurements.

Relations of the Quantities

The principal difference between the phenomena of radio (high) and audio (low) frequency is the importance of capacity and inductance at radio frequency as compared with the predominance of resistance in audio frequency phenomena. Nevertheless, resistance is the measure of power consumption, since any dissipation or loss of electrical power is expressible in terms of a resistance. Furthermore, resistances change rapidly with frequency at radio frequencies. The change can be calculated for certain very simple cases, but in most practical cases the resistance can be determined only by measurement. Thus the measurement of resistance at radio frequencies is a necessary and important operation.

Certain related quantities also express power dissipation. The results of a measurement can be expressed in terms of any of these quantities when their relations are clearly established.

In a simple series circuit, Figure 1, when the emf. is a sustained sine wave, a maximum of current is obtained when the inductive reactance is just equal to the capacitive reactance, i. e., when the equation

$$I=\frac{E}{\sqrt{R^2+\left(\omega L-\frac{1}{\omega C}\right)^2}}$$

* Received by the Editor, April 17, 1918.

reduces to $I_r = \frac{E}{R}$. This condition of resonance obtains when

$$\omega L - \frac{1}{\omega C} = 0$$

or

$$\omega = \frac{1}{\sqrt{CL}} \tag{1}$$

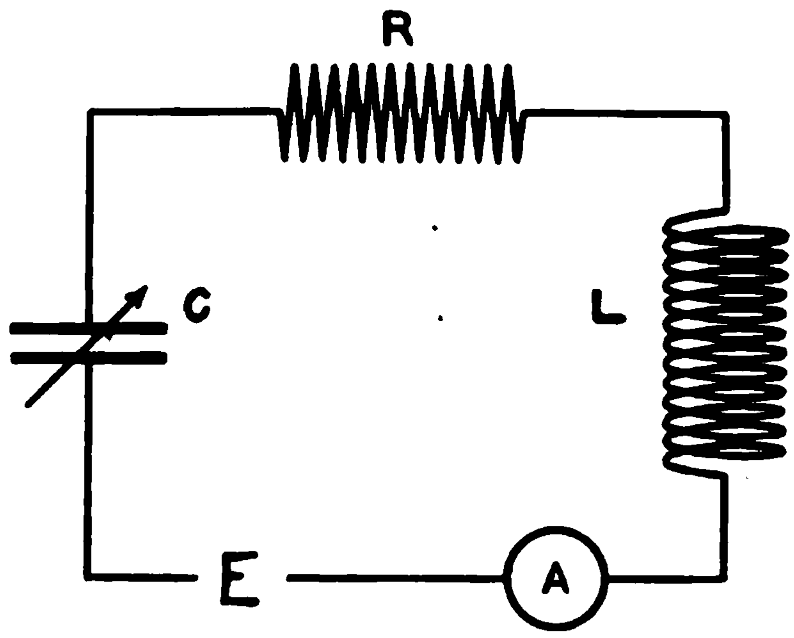

FIGURE 1

For any variation of either ω, L, or C, from this condition, as at A for the curves of Figure 2, the current is smaller than the value for which this relation holds.

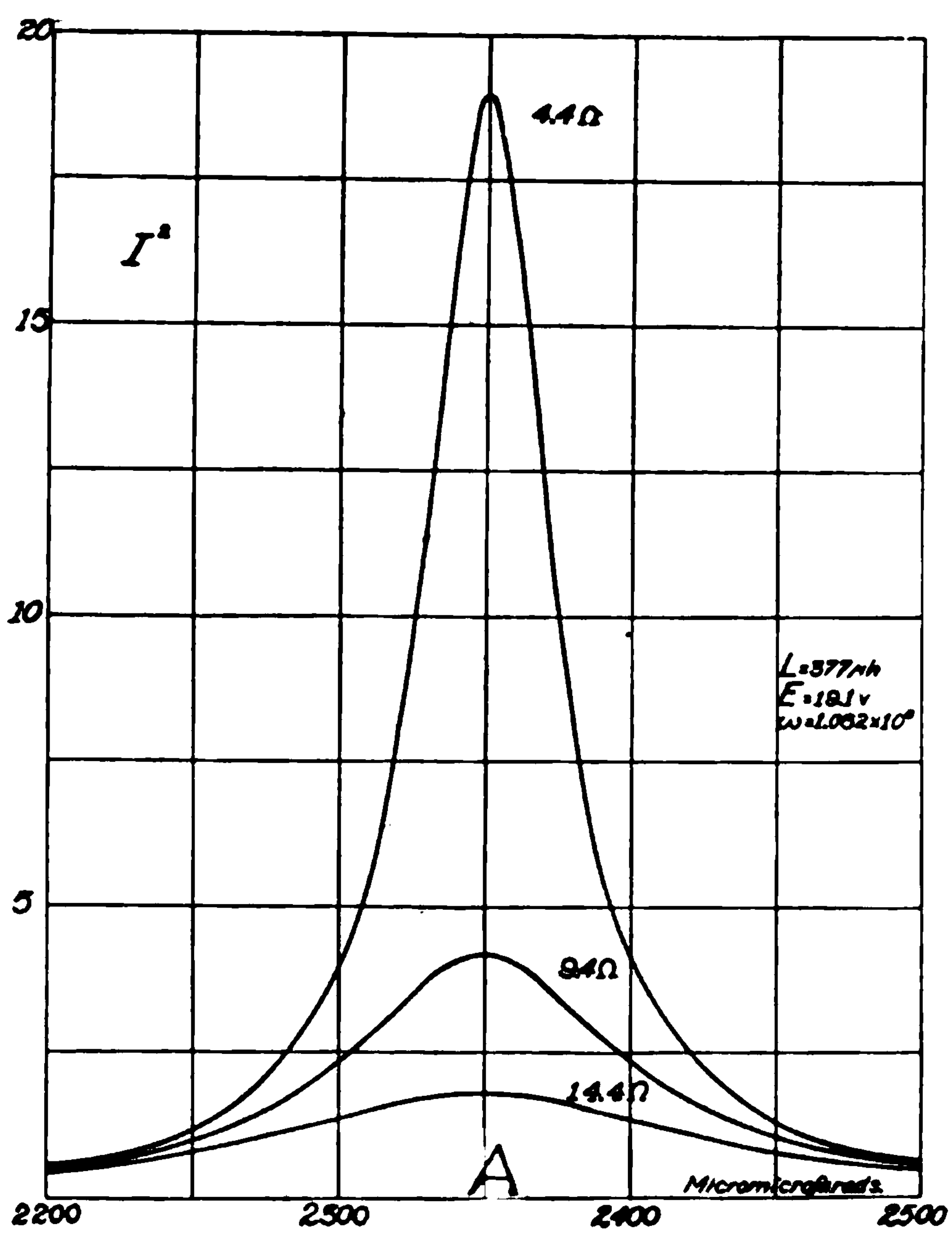

FIGURE 2

The phase angle of the circuit is zero at resonance as shown in Figure 3. The expression, phase angle of the circuit, means the same thing as phase angle of the current in the circuit.

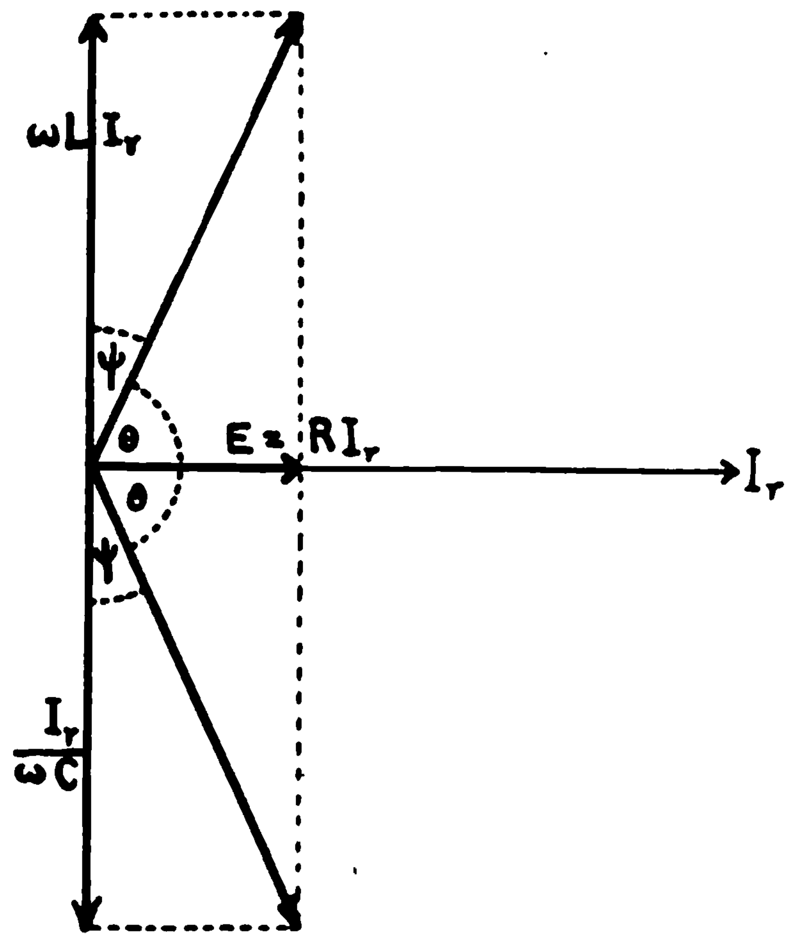

FIGURE 3

The phase angle θ of the coil, considering the resistance R to be associated with the coil L, is equal to the phase angle θ of the condenser, considering the resistance to be associated with the condenser. The complement of the phase angle θ is the phase difference ψ. From Figure 3,

$$\tan\psi = \frac{R}{\omega L} = R\,\omega C$$

When R and ψ are small, as usually in radio circuits, the tangent equals the angle and

$$\psi = \frac{R}{\omega L} = R\,\omega C \tag{2}$$

Phase difference is thus a ratio of resistance to reactance. It follows also that $\psi = \sin\psi = \cos\theta$, or phase difference is equal to power factor. The same relation is brought out by multiplying in (2) by I^2.

$$\psi = \frac{R\,I^2}{\omega L\,I^2}$$

or

$$\psi = \frac{R\,I^2}{\dfrac{I^2}{\omega C}}$$

= ratio of power dissipated to power flowing.

Thus

$$\psi = \textit{power factor.} \tag{3}$$

Another quantity of importance in connection with the expression of power dissipation is the sharpness of resonance. This is the quantity which measures the fractional change in current for a given fractional change in either the capacitive or inductive reactance from its value at resonance. (Practically the same quantity has been known as persistency, selectivity, and resonance ratio.) It may be defined in mathematical terms for a variation of C by the following ratio

$$S = \frac{\sqrt{\dfrac{I_r^2 - I_1^2}{I_1^2}}}{\dfrac{\pm(C_r - C)}{C}}$$

where the subscript "$_r$" denotes value at resonance, and I_1 is some value of current corresponding to a capacity C which differs from the resonance value. The numerator of this expression is somewhat arbitrarily taken to be the square root of the fractional change in the current-square instead of taking directly the fractional change of current. This is done partly because of the convenience in actual use of this expression (since the deflections of the usual detecting instruments are proportional to the square of the current), and also because of mathematical convenience. It is easily seen qualitatively that sharpness of resonance is large when phase difference is small; as when R is small compared with ωL, a given fractional change in ωL changes the impedance, and therefore the current, by a relatively large fractional amount. Quantitatively, from the relations,

$I_1^2 = \dfrac{E^2}{R^2 + \left(\omega L - \dfrac{1}{\omega C}\right)^2}$, $I_r^2 = \dfrac{E^2}{R^2}$, and $\omega L = \dfrac{1}{\omega C_r}$, it follows at once that

$$S = \frac{\sqrt{\dfrac{I_r^2 - I_1^2}{I_1^2}}}{\dfrac{\pm(C_r - C)}{C}} = \frac{1}{R\,\omega C_r} = \frac{\omega L}{R} = \frac{1}{\psi} \tag{4}$$

= ratio of power flowing to power dissipated. Thus,

$$\text{Sharpness of resonance} = \text{reciprocal of power factor.} \tag{5}$$

Another related quantity is the logarithmic decrement. For free oscillation in a simple circuit, the value of the decrement

or napierian logarithm of the ratio of two successive current maxima in the same direction is readily shown to be

$$\delta = \pi R \sqrt{\frac{C}{L}} = \frac{\pi R}{\omega L} = \pi R \omega C \tag{6}$$

Comparing with (2) and (4),

$$\delta = \pi \psi = \frac{\pi}{S} \tag{7}$$

That is, altho defined originally in connection with damped oscillations, the decrement is a constant of the circuit and can be dealt with in the same way as resistance and the other energy-determining quantities. It is similarly possible to speak of the decrement of a part of a circuit in the same manner as phase difference and resistance.

The decrement is definable in terms of an energy ratio. Thus

$$\delta = \pi \psi = \frac{\pi R}{\omega L} = \frac{\pi R I^2}{\omega L I^2} = \frac{\frac{\pi R I^2}{2 \pi f}}{L I^2} = \frac{1}{2} \frac{\frac{R I^2}{f}}{L I^2}$$

For sustained sinusoidal oscillations, $\frac{RI^2}{f}$ = energy dissipated per cycle, and $L I^2 = \frac{1}{2} L (2 I^2) = \frac{1}{2} L I_o^2$ = magnetic energy associated with the current at the maximum of the cycle. It follows that the decrement is one-half the ratio of the energy dissipated per cycle to the energy associated with the current at the maximum of the cycle. This relation holding for each cycle, it holds for the average of all the cycles.

That the same definition of decrement applies to the natural damped oscillations of a circuit may be shown as follows. Suppose the circuit to be set oscillating, so that a train of natural oscillations takes place, N times per second, and that the energy of each train is practically all dissipated before the next one begins. N is the group frequency or number of complete trains of oscillations per second. The energy dissipated during a train of waves equals the energy input at the beginning of each train $= \frac{1}{2} L I_o^2$, where I_o is the first maximum of current. The average energy dissipated per cycle must equal the energy dissipated during a train of waves divided by the number of cycles in a train. The number of cycles in a train is the ratio of the frequency of oscillations f to the group frequency N. Therefore, the average energy dissipated per cycle $= \frac{\frac{1}{2} L I_o^2}{\frac{f}{N}} = \frac{N L I_o^2}{2f}$. The average energy associated with the

current at the maximum of each cycle may be shown to be $L I^2$, where I^2 = root-mean-square current, just as in the case of undamped currents, provided the decrement is not large. Applying the energy-ratio definition of decrement,

$$\delta = \frac{1}{2} \frac{\frac{N L I_o^2}{2f}}{L I^2} = \frac{N I_o^2}{4 f I^2} \tag{8}$$

This checks the familiar equation for the root-mean-square value of natural oscillations,

$$I^2 = \frac{N}{4 f \delta} I_o^2 = \frac{N}{4 a} I_o^2 \tag{9}$$

This definition of decrement, one-half the ratio of the average energy dissipated per cycle to the average energy associated with the current at the maximum of each cycle, is a valuable conception. It has here been shown to apply to both undamped oscillations and to natural damped oscillations.

It is thus evident that resistance, phase difference, sharpness of resonance, and decrement are all constants of a circuit expressing energy dissipation or power factor. Their relations are given in equations (2) to (7). Using these relations, a measurement of any one of them can be made to give all the others. Since resistance is the simplest quantity, specific consideration is given in the following to resistance measurement. Nevertheless in some cases one of the other quantities is the more convenient or more useful one in terms of which to express the results of measurement.

Methods of Measurement

General—There is considerable difficulty in attaining high accuracy in measurements at radio frequencies. Much of this is due to the fact that the quantities to be measured or upon which the measurement depends are generally small and sometimes not definitely localized in the circuits. Thus the inductances and capacities used in the measuring circuits are so small that the effect upon these quantities of lead wires, indicating instruments, surroundings, etc., must be carefully considered. The capacity of the inductance coil and sometimes even the inductance within the condenser are of importance. In order to minimize these various effects, it is generally best to use measuring circuits and methods which are the least complicated. On this account simple circuits and substitution

methods in which the determination depends upon deflections are usually used in preference to more complicated methods.

In addition to the uncertainty or the distributed character of some of the quantities to be measured, there are other limitations upon the accuracy of radio measurements. The usual ones are the variation with frequency of current distribution, of inductance, resistance, and so on, and the difficulty of supplying radio frequency current of sufficient constancy. The latter limitation is entirely overcome by the use of the electron tubes as a source of current, but is troublesome when a buzzer, spark, or arc is used. As to the other difficulty: the variations of inductance, and so on, with frequency, while these variations have a profound effect, they are generally subject to control; the quantities have definite values at a particular frequency under definite conditions, and their effect can usually be determined by calculation or measurement.

It is not always possible to determine the effects of the capacities of accessory apparatus and surroundings, nor to eliminate them, and thus they remain the principal limitation upon the accuracy of measurements. These stray capacities include the capacities of leads, instrument cases, table tops, walls, and the observer. They may not only be indeterminate but may vary in an irregular manner.

Classification of Methods

On account of the requirement of simplicity in radio measurements, the methods available are quite different, and are fewer in number, than in the case of audio-frequency or direct-current measurements.

The methods of measuring radio-frequency resistance may be roughly classed as:

(1) Calorimeter method.
(2) Substitution method.
(3) Resistance-variation method.
(4) Reactance-variation method.

The fourth has frequently been called the "decrement method," but it is primarily a method of measuring resistance rather than decrement, exactly as the resistance variation method is. Either may be used to measure the decrement of a wave under certain conditions, and in fact the results of resistance measurement by any method may be expressed in terms of decrement.

All four methods may be used with either damped or undamped waves, tho in some of them the calculations are different in the two cases. They are all deflection methods, in the sense of depending upon the deflections of some form of radio-frequency ammeter. In the first and second, however, it is only necessary to adjust two deflections to approximate equality, while in the third and fourth the deflections may have any magnitude.

Calorimeter Method

This method may be used to measure the resistance either of a part or the whole of a circuit. The circuit or coil or other apparatus, the resistance of which is desired, is placed in some form of calorimeter, which may be a simple air chamber, an oil bath, or other suitable form. The current is measured by an accurate radio-frequency ammeter, and the resistance R_x is calculated from the observed current I and the power, or rate of heat production, P

$$P = R_x I^2 \tag{10}$$

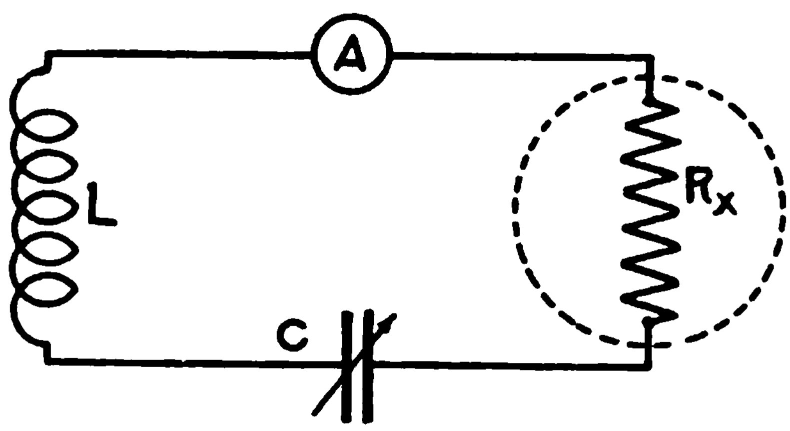

Figure 4

While P might be measured calorimetrically, in practice it is always measured electrically by an auxiliary observation in terms of audio-frequency or direct current Thus it is only necessary to observe the temperature of the calorimeter in any arbitrary units when the radio-frequency current flows, and then cause audio-frequency current to flow in the circuit, adjusting its value until the temperature becomes the same as before. Denoting by the subscript "$_o$" the audio-frequency values

$$P_o = R_o I_o^2$$

$$\frac{P}{P_o} = \frac{R_x I^2}{R_o I_o^2}$$

$$\frac{R_x}{R_o} = \frac{P}{P_o} \frac{I_o^2}{I^2}$$

For $P = P_o$, $$R_x = R_o \frac{I_o^2}{I^2} \tag{11}$$

From the known audio frequency value of the resistance, therefore, and the observed currents, the resistance is obtained.

The radio and audio frequency observations are sometimes made simultaneously, using another resistance of the same magnitude as that of the apparatus, the resistance of which is desired, placed in another calorimeter as nearly identical with the first as possible. High (radio) frequency current is passed thru one, low (audio) frequency thru the other, and the calorimeters kept at equal temperatures by means of some such device as a differential air thermometer or differential thermoelement. To compensate for inequalities in the two sets of apparatus, the radio and audio frequency currents are interchanged. This method may be found more convenient in some circumstances, but the extra complication of apparatus is usually not worth while, and the value of the measurement depends upon the accurate observation of the radio frequency current I, just as the simpler method does.

The calorimeter method, while capable of high accuracy, is slow and less convenient than some of the other methods. It has been used by a number of experimenters to measure the resistance of wires and coils.

Substitution Method

This method is applicable only to a portion of a circuit. Suppose that in Figure 4 the coil L is loosely coupled to a source of oscillations The capacity C is varied until resonance is obtained, and the current in the ammeter is read A resistance standard is then substituted for the apparatus R_x and varied until the same current is indicated at resonance. If the substitution has changed the total inductance or capacity of the circuit, the retuning to resonance introduces no error when undamped or slightly damped electromotive force is supplied, provided the change of condenser setting introduces either a negligible or known resistance change. In the case of a rather highly damped source, however, the method can only be used when the resistance substitution does not change the inductance or capacity of the circuit. The unknown R_x is equal to the standard resistance inserted, *provided* the electromotive force acting in the circuit has not been changed by the substitution of the standard for R: this condition is discussed below.

The resistance standards usually used are not continuously

variable, and hence the standard used may give a deflection of the ammeter somewhat different from the original deflection. To determine the resistance in this case, three deflections are required, all at resonance. In one application, the apparatus of unknown resistance R_x is inserted and the current I_x observed; then a similar apparatus of known resistance R_n is substituted for it and the current I_n observed; and finally a known resistance R_1 is added and the current I_1 observed. The relations between these quantities and the electromotive force involve the unknown but constant resistance of the remainder of the circuit R, thus,

$$R_x + R = \frac{E}{I_x}$$

$$R_n + R = \frac{E}{I_n}$$

$$R_1 + R_n + R = \frac{E}{I_1}$$

from which

$$R_x - R_n = R_1 \frac{\frac{I_n}{I_x} - 1}{\frac{I_n}{I_1} - 1} \tag{12}$$

This method is closely related to the resistance variation method; see formula (13) below.

The substitution method is very convenient and rapid and is suitable for measurements upon antennas, spark gaps, etc., and for rough measurements of resistances of condensers and coil. In radio laboratory work, however, using delicate instruments and with loose coupling to the source of oscillations, it is found that it is not a highly accurate method, except for measuring small changes in resistance of a circuit. The reason for this is that there are other electromotive forces acting in the circuit than that purposely introduced by the coupling coil, viz., emf.'s electrostatically induced between various parts of the circuit. When the apparatus under measurement is removed from the circuit, these emf.'s are changed, and there is no certainty that when the current is made the same the resistance has its former value. Something of the same difficulty enters into the question of grounding the circuit in the following method, as discussed below.

Resistance Variation Method

This method measures primarily the effective resistance of the whole circuit, including that due to condenser losses and

to radiation. The principle may be readily understood from the diagram of the simple circuit, Figure 5.

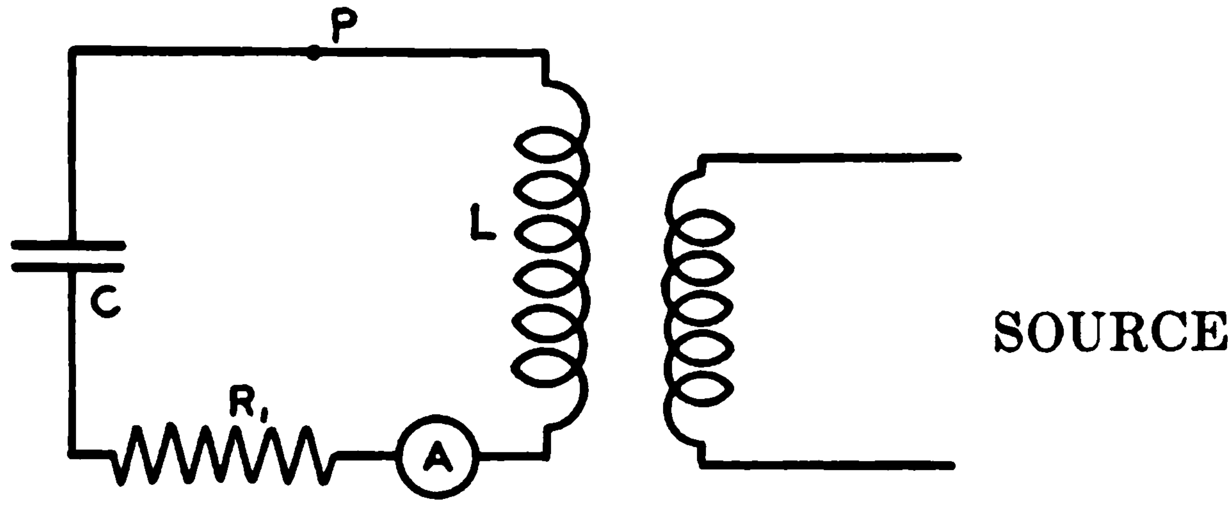

FIGURE 5

If the resistance of some particular piece of apparatus, inserted at P for example, is to be found, the resistance of the circuit is measured with it in circuit and then re-measured in the same way with it removed or replaced by a similar apparatus of known resistance; and the resistance of the apparatus is obtained by simple subtraction.

The results of a measurement may be expressed in terms of ψ, S, or δ by equations (2) to (7) above. The method, however, is particularly convenient where resistance is the actual quantity the value of which is wanted.

The measurement is made by observing the current I in the ammeter A when the resistance R_1 has its zero or minimum value, then inserting some resistance R_1 and observing the current I_1. Let R denote the resistance of the circuit without added resistance. Suppose that a sine-wave electromotive force E is introduced into the circuit by induction in the coil L from a source of undamped waves, and that the two observations are made at resonance. For the condition of resonance,

$$I=\frac{E}{R}$$

$$I_1=\frac{E}{R+R_1}$$

from which the resistance of the circuit is given by

$$R=R_1\frac{I_1}{I-I_1}=\frac{R_1}{\frac{I}{I_1}-1} \tag{13}$$

The same method can be employed using damped instead

of continuous waves, and can even be used when the current is supplied by impulse excitation, but the equations are different; see (59) and (16) below. When the damping of the supplied emf. is very small, equation (13) applies.

PRECAUTIONS

A limitation on the accuracy of the measurement is the existence of the emf.'s electrostatically induced that were mentioned above. In the deduction of (13) it is assumed that E remains constant. The virtue of this method is that these emf.'s may be kept substantially constant during the measurement of resistance of the circuit. They will invariably be altered by the insertion of the apparatus, the resistance of which is desired, but the resistance of the circuit is measured accurately in the two cases and the difference of the two measurements gives the resistance sought. In order to keep these stray electromotive forces unchanged when R_1 is in and when it is out of circuit, particular attention must be paid to the grounding of the circuit. The shield of the condenser and the ammeter (particularly if it is a thermocouple with galvanometer) have considerable capacity to ground and are near ground potential. A ground wire, if used, must be connected either to the condenser shield or to one side of the ammeter. If connected to the high-potential side of the inductance coil, absurd results will be obtained. The resistance R_1 also must be inserted at a place of low potential, preferably between the condenser and ammeter.

USE OF THERMOCOUPLE

Another necessary precaution is to keep the coupling between source and measuring circuit so loose that there is no reaction. This necessitates the use of a sensitive device for current measurement. As regularly carried out at the Bureau of Standards, in the resistance variation method, a pliotron is used as a source of undamped emf., and current is measured with a thermocouple in series in the measuring circuit. The currents corresponding to given deflections of the thermocouple galvanometer are obtained from a calibration curve, or from the law $d \propto I^2$, where d = deflection, if the instrument follows this law sufficiently closely. When the deflections follow this law, equation (13) becomes

$$R = \frac{R_1}{\sqrt{\frac{d}{d_1}} - 1} \tag{14}$$

Several values of resistance R_1 are usually inserted in the circuit and the corresponding deflections obtained; the resulting values of R are averaged.

When the thermocouple follows the square law accurately, the quarter deflection method may be used, which eliminates all calculation. When the deflection d_1 is $\frac{d}{4}$, equation (14) becomes

$$R = R_1 \tag{15}$$

This method requires a variable resistance standard such that R_1 can be varied continuously in order to make d_1 just equal to $\frac{d}{4}$. Practically the same method is used if the resistance is varied by small steps, as in a resistance box, and interpolating between two settings of R_1.

Use of Impulse Excitation

The procedure for the resistance variation method is the same when the current is damped as when undamped. When the circuit is supplied by impulse excitation, so that free oscillations are produced, the theory of the measurement is very simple. The current being I when the resistance is R, and I_1 when the resistance R_1 is added, the power dissipated in the circuit must be the same in the two cases because the condenser in the circuit is charged to the same voltage by each impulse which is impressed upon it, and there is assumed to be no current in the primary after each impulse.
Therefore

$$R I^2 = (R + R_1) I_1^2$$

whence,

$$R = R_1 \frac{I_1^2}{I^2 - I_1^2} \tag{16}$$

It is difficult to obtain high accuracy by the method in practice because of the difficulty of obtaining pure impulse excitation.

The method is specially convenient when an instrument is used in which the deflection d is proportional to the current squared. Then (16) becomes

$$R = R_1 \frac{d_1}{d - d_1} \tag{17}$$

This is still further simplified if the resistance R_1 is adjustable

so that d_1 can be made equal to one-half d. The equation then reduces to

$$R = R_1 \tag{18}$$

This is commonly known as the half-deflection method.

Use of Damped Excitation

The resistance variation method has already been shown to be usable with either undamped or free oscillations. It can also be used when the supplied emf. is damped so that both forced and free oscillations exist in the circuit. The equations (56) to (65) below show how the decrement of a circuit is obtained from such measurements. Resistance is then readily calculated by equation (6). As already stated, when the damping of the supplied emf. is extremely small, equation (13) applies. The decrement of the supplied emf. may itself be obtained by such measurements whether the emf. be due to a nearby circuit or to a wave travelling thru space.

Application of Method

This method is used in precision measurements upon condensers, coils, wavemeters, etc. The accurate measurement of resistance of a wavemeter circuit is of particular importance because the wavemeter is frequently used to measure the resistance, phase difference, or decrement of other apparatus. It is the calibration of a resistance-measuring standard.

The resistance of a wavemeter is not a single constant value. It varies with frequency and with the detecting or other apparatus connected to the wavemeter circuit. Usually both the resistance and the decrement of the circuit vary with the condenser setting. It is usually desirable to express either resistance, sharpness of resonance, or decrement in the form of curves for the several wavemeter coils, each for a particular detecting apparatus or other condition.

When a pliotron, arc, or other source of undamped wave is used, formula (13) above is used. When the current-measuring device is a current-square meter, thermocouple or crystal detector with galvanometer, or other apparatus which is so calibrated that deflections are accurately proportional to the square of the current, and when in addition a continuously variable resistance standard is used, the quarter-deflection method may be employed eliminating all calculation.

When a buzzer or other source is used, arranged to give impulse excitation, equation (16) above gives the resistance.

When the current indicator is calibrated in terms of the square of the current and the resistance standard is continuously variable, the measurement is conveniently made by the half-deflection method.

Reactance Variation Method

This has been called the decrement method, a name which is no more applicable to this than to the other methods of resistance measurement since all measure decrement in the same sense that this does. That the method primarily measures resistance rather than decrement is seen from the fact that in its simple and most accurate form it utilizes undamped current, which has no decrement.

The method is analogous to the resistance variation method, two observations being taken. The current I_r in the ammeter (Figure 6) is measured at resonance, the reactance is then varied and the new current I_1 is observed. The total resistance of the circuit R (including that due to condenser losses, radiation, etc.) is calculated from these two observations. The reactance may be varied by changing either the capacity, the inductance, or the frequency, the emf. being maintained constant. The reactance is zero at resonance and it is changed to some value X_1 for the other observation. With undamped emf. E, the currents are given by

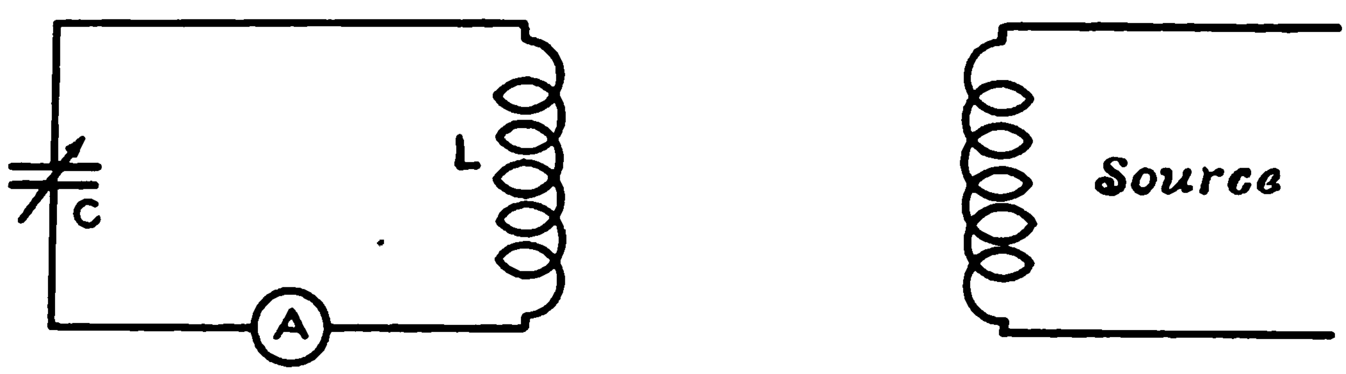

Figure 6

$$I_r^2 = \frac{E^2}{R^2}$$

$$I_1^2 = \frac{E^2}{R^2 + X_1^2}$$

From these it follows that

$$R = X_1 \sqrt{\frac{I_1^2}{I_r^2 - I_1^2}} \tag{19}$$

This has a similarity to $R=R_1\frac{I_1}{I-I_1}$, the equation (13) for the resistance-variation method. It is also interesting that when the reactance is varied by such an amount as to make the quantity under the radical sign equal to unity, the equation reduces to

$$R=X_1 \tag{20}$$

This is similar to $R=R_1$, which is the equation for the quarter-deflection and half-deflection resistance-variation methods.

Resistance Measurement

When the reactance is varied by changing the setting of a variable condenser,

$$X_1=\pm\left(\frac{1}{\omega C}-\frac{1}{\omega C_r}\right)$$

and the equation (19) becomes

$$R=\frac{\pm(C_r-C)}{\omega C_r C}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{21}$$

For variation of the inductance, (19) becomes

$$R=\pm\,\omega\,(L-L_r)\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{22}$$

and for variation of the frequency

$$R=\frac{\pm L\,(\omega^2-\omega_r^2)}{\omega}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{23}$$

This equation is equivalent to

$$R=\frac{\pm\,6\,\pi\times10^8L\,(\lambda_r^2-\lambda^2)}{\lambda\,\lambda_r^2}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{24}$$

for λ in meters, R in ohms, and L in henrys.

It must be noted that variation of the frequency or wave length requires some alteration in the source of emf., and the greatest care is necessary to insure that the condition of constant emf. is fulfilled. This is discussed below in connection with equations (30) and (31).

In the use of equation (22), some error is introduced into the measurement if the variable inductor is also used as the coupling to the source, on account of the variation thus introduced into the E supplied. The per cent. error, however, is usually not more than the per cent. variation of L.

A convenient method which differs slightly from those just

described is to observe two values of the reactance both corresponding to the same current I_1 on the two sides of the resonant value I_r. For observation in this manner of two capacity values C_1 and C_2,

$$R=\frac{1}{2\omega}\frac{C_2-C_1}{C_2C_1}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{25}$$

The simple derivation here given for these formulas is much shorter than the usual treatments, and at the same time is more comprehensive. These formulas are all rigorous, involving no approximations, provided the applied emf. is undamped. They also apply for damped emf. when the damping is negligibly small.

It is customary to reduce the labor of computation by varying the reactance by such an amount that $I_1^2=\frac{1}{2}I_r^2$, making the quantity under the radical sign equal to unity, so that formulas (21) to (25) are much simplified.

Measurements of Phase Difference, Sharpness of Resonance, and Decrement

Measurements by the reactance variation method are very conveniently expressed in terms of phase difference, sharpness of resonance, and decrement. The formulas are in fact simpler for any of these quantities than for resistance. Thus, utilizing equations (2) to (7) it is readily found that (21) is equivalent to:

$$\psi=\frac{\pm(C_r-C)}{C}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{26}$$

$$S=\frac{C}{\pm(C_r-C)}\sqrt{\frac{I_r^2-I_1^2}{I_1^2}} \tag{27}$$

$$\delta=\pi\frac{\pm(C_r-C)}{C}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{28}$$

Equation (27) is identical with (4) above, thus suggesting that the definition of sharpness of resonance itself contains inherently this method of measurement. The equations corresponding to (22) to (25) are obtained for ψ, S, and δ, in the same manner as (26) to (28). Those for phase difference, expressed in radians, are

$$\psi=\frac{\pm(L-L_r)}{L_r}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{29}$$

$$\psi=\frac{\pm(\omega^2-\omega_r^2)}{\omega\,\omega_r}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \tag{30}$$

$$\psi = \frac{\pm(\lambda_r^2 - \lambda^2)}{\lambda \lambda_r} \sqrt{\frac{I_1^2}{I_r^2 - I_1^2}} \tag{31}$$

$$\psi = \frac{C_2 - C_1}{C_2 + C_1} \sqrt{\frac{I_1^2}{I_r^2 - I_1^2}} \tag{32}$$

Phase difference is a particularly convenient constant in terms of which to express the results of measurements upon condensers, since the phase difference of most condensers is usually a constant with respect to frequency at radio frequencies. These formulas are rigorous provided ψ is small, as it usually is in radio circuits, when the emf. is sustained, and hold also for damped emf. when the damping is negligibly small. The use of the method when the applied emf. has a moderate damping is discussed in the last section of this paper.

A convenient way to utilize the method indicated in (30) and (31) is to vary the wave length by means of a variable condenser or inductor in the source circuit. An incorrect formula has sometimes been given for decrement measurement by this method. The following are rigorous:

$$\psi = \frac{\pm(C_r' - C')}{\sqrt{C_r' C'}} \sqrt{\frac{I^2_1}{I_r^2 - I_1^2}}$$

$$\delta = \pi \frac{\pm(C_r' - C')}{\sqrt{C_r' C'}} \sqrt{\frac{I_1^2}{I_r^2 - I_1^2}}$$

where the capacities are those of the condenser in the source circuit. This method must be used with great caution because constancy of E, the applied emf., is required. The source circuit is necessarily disturbed by the variation of its condenser setting; when the variation is small, and a pliotron is used as the source, the current I' in the source circuit may not be appreciably changed. It is desirable to use a sensitive indicating instrument and actually observe I'. Constancy of I', however, would not mean that the emf. acting on the measuring circuit was constant, for $E = \omega M I'$, and thus E varies by the amount of the ω variation. The per cent. error in the resulting value of ψ or δ equals the per cent. change of C', when the method is made by the familiar procedure which reduces the current ratio under the radical to unity.

Direct-reading Phasemeters and Decremeters

A phasemeter as used in radio work is a wavemeter conveniently arranged for measurements of phase difference. A decremeter is a wavemeter similarly arranged for measurements

of decrement. While, of course, resistance and sharpness of resonance can be calculated from measured values obtained by either of these instruments, the principal application of a phasemeter is in measurements of phase difference of condensers and of dielectric materials and the principal use of a decremeter is in the measurement of the decrement of a wave. The forms of these instruments usually employed make use of the reactance-variation method. Any such instrument may be used either as a phasemeter or a decremeter by merely changing the instrument scale by a constant factor. While decremeters have been more commonly used, the phasemeter is a somewhat more direct application of the underlying theory. In the development of the theory of the instrument, undamped (sustained) sine-wave emf. is assumed.

Determination of the Scale of a Phasemeter or Decremeter

Any wavemeter, the circuit of which includes some form of ammeter, may be fitted with a special scale from which phase difference or decrement may be read directly. The procedure for a wavemeter having any sort of variable condenser is given here.

The usual use of the reactance-variation method is in accordance with equation (32), the currents being adjusted so as to make the quantity under the radical unity. That is, the current-square meter is first observed at resonance, the variable condenser is reset to a value C_1 on one side of resonance such that the current-square is reduced to one-half, and then set to another value C_2 on the other side of resonance giving the same current-square. The phase difference is calculated by

$$\psi = \frac{C_2 - C_1}{C_2 + C_1} \tag{33}$$

A certain value of phase difference, therefore, corresponds to that displacement of the condenser's moving plates which varies the capacity by the amount $(C_2 - C_1)$. The displacement for a given phase difference will, in general, be different for different values of C, the total capacity in the circuit. At each point of the condenser scale, therefore, any displacement of the moving plates which changes the square of current from $\frac{1}{2} I_r^2$ on one side of resonance to the same value on the other side means a certain value of ψ.

A special scale may, therefore, be attached to any variable

condenser, with graduations upon it and so marked that the difference between the two settings on the two sides of resonance is equal to the phase difference. The spacing of the graduations at different parts of the scale depends upon the relation between capacity and displacement of the moving plates. When this relation is known, the scale can be predetermined. A scale may, therefore, be fitted to any condenser, from which phase difference may be read directly, provided the capacity of the circuit is known for all settings of the condenser. The scale may be attached either to the moving plate system or to the fixed condenser top. It is usually convenient to attach it to the unused half of the dial opposite the capacity scale.

The scale for such an instrument is determined as follows. When the change of capacity setting is small, as usually in radio work, (33) may be written

$$\psi = \frac{dC}{2C} \tag{34}$$

Letting s denote readings on the required scale, the value of ψ is the difference of two s readings, or

$$\psi = ds$$

$$ds = \frac{dC}{2C}$$

The readings of the scale are then given by

$$s = \int_{C}^{C_a} \frac{dC}{2C}$$

$$s = \tfrac{1}{2}\,(\log_e C_a - \log_e C) \tag{35}$$

C_a is the arbitrary capacity chosen as the zero point of the scale. Thus the scale can begin anywhere. Such a scale gives ψ in radians.

A wavemeter in which the inductance is variable and the capacity fixed is also convertible into a phasemeter in similar manner. The instrument is operated in just the same way, and the equation, corresponding to (33), is

$$\psi = \frac{L_2 - L_1}{L_2 + L_1} \tag{36}$$

and the direct-reading phasemeter scale is given by

$$s = \tfrac{1}{2}\,(\log_e L_a - \log_e L) \tag{37}$$

where L_a is the arbitrary inductance chosen as the zero point of the scale.

A direct-reading decremeter is made in precisely the same

way as a phasemeter. The decrement is π times the phase difference in radians, hence the equations for a decrement scale on a wavemeter with a variable condenser or variable inductor are respectively

$$s = \frac{\pi}{2}\left(\log_e C_a - \log_e C\right) \tag{38}$$

$$s = \frac{\pi}{2}\left(\log_e L_a - \log_e L\right) \tag{39}$$

The phase difference or decrement measured by such an instrument, using undamped sine-wave emf., is the phase difference or decrement of the measuring circuit itself. Its application to measuring the decrement of a wave is explained in the last section below. When the instrument is used as the measuring circuit with undamped emf., the variable condenser or inductor must be one having zero effective resistance or in which the resistance for each setting and each wave length is accurately known, in order that the resistance or ψ or δ of other apparatus connected in the circuit may be obtained. When a variable condenser is used as the phasemeter it is thus convenient for measurements upon coils, and when a variable inductor is the phasemeter it is a convenient means for measurements upon the R or ψ or δ of any condenser connected to it.

The direct-reading phasemeter or decremeter may also be used in the source circuit, to vary the ω or λ supplied to the measuring circuit, as described in connection with equations (30) and (31) above. In this use it is not necessary to make correction for the resistance of the phasemeter or decremeter itself, whether it be variable condenser or variable inductor, as it has no effect upon the measuring circuit, except insofar as it may effect the value of ω, which would be a second-order effect. This use of the direct-reading phasemeter is now being exhaustively studied by Messrs. G. C. Southworth and J. L. Preston at the Bureau of Standards.

Simple Direct-reading Phasemeter or Decremeter

It is particularly easy to make a phasemeter or decremeter out of a condenser with semi-circular plates. Such condensers follow closely the linear law,

$$C = a\beta + C_o \tag{40}$$

where β is the angle of rotation of the moving plates and a and C_o are constants. It can be shown that the phase difference scale applicable to such a condenser is one in which the graduations vary as the logarithm of the angle of rotation. Furthermore,

the same scale applies to all condensers of this type. This scale has been calculated and is given in Figure 7 for values of phase difference in degrees.

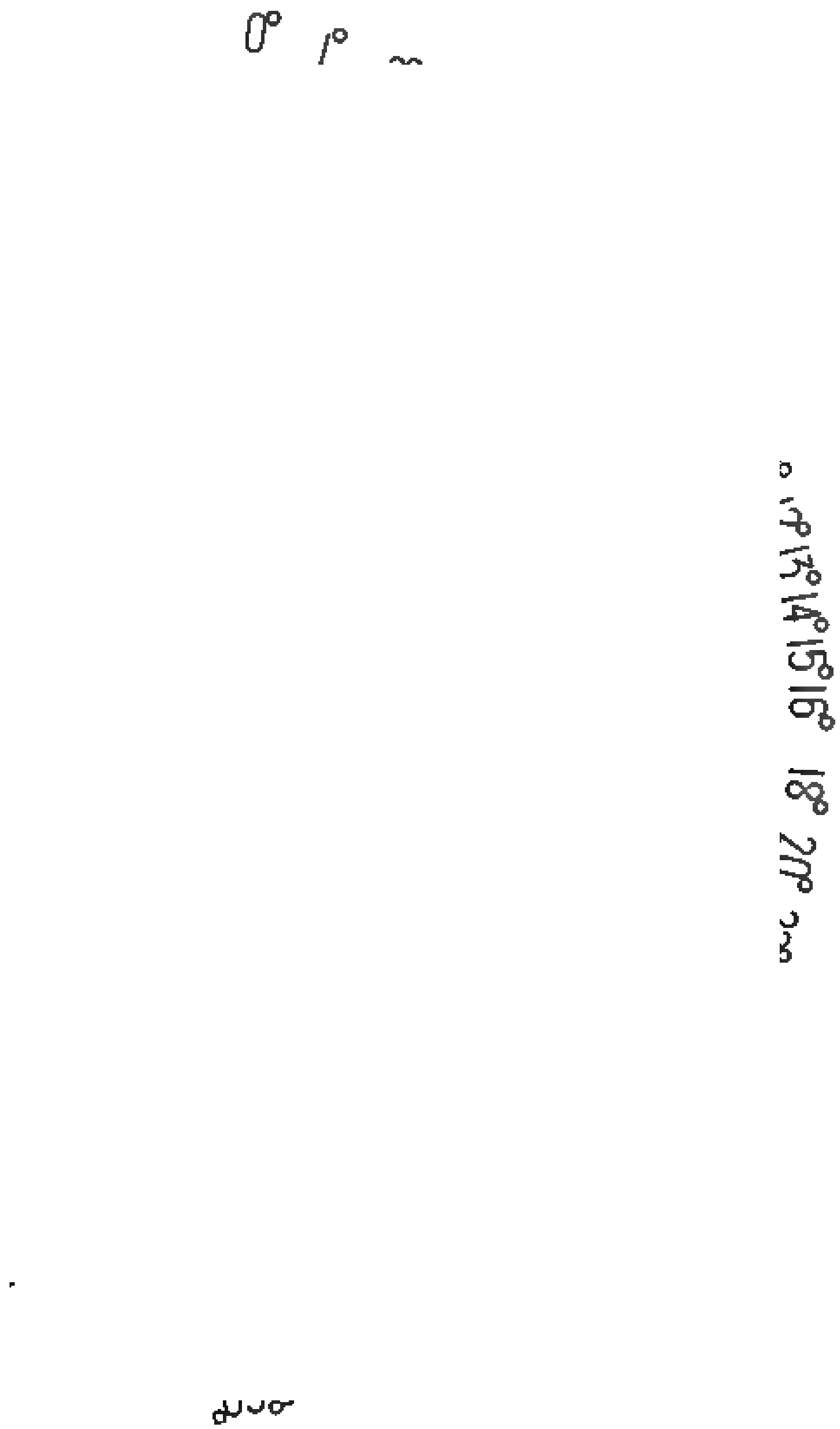

FIGURE 7

The scale was calculated in the following manner. Inserting equation (40) in (35)

$$s = \tfrac{1}{2}\left[\log_e (a\,\beta_o + C_o) - \log_e (a\,\beta + C_o)\right] \tag{41}$$

Let $C_o = a\,\beta_o$,

$$s = \frac{2.303}{2}\left[log_{10}(\beta_a+\beta_o) - log_{10}(\beta+\beta_o)\right] \tag{42}$$

For $C_o = \beta_o = 0$,

$$s = 1.151\,(log_{10}\,\beta_a - log_{10}\,\beta) \tag{43}$$

$$log_{10}\,\beta = log_{10}\,\beta_a - \frac{s}{1.151}$$

For s expressed in degrees rather than radians, and for $\beta_a = 180°$, the angular separation in degrees on the ψ scale is given by

$$log_{10}\,\beta = log_{10}\,180 - \frac{s}{57.3\,(1.151)} \tag{44}$$

This scale may be used as it stands on any variable condenser with semi-circular plates, regardless of the kind of capacity scale on the condenser or even if the condenser has no scale whatever on it. The phase difference scale may, if desired, be cut out and trimmed at such a radius as to fit the dial and then affixed to the condenser, with its zero point approximately in coincidence with the graduation which corresponds to maximum capacity. This usually puts it on the unused half of the dial opposite the capacity scale. If the figures are trimmed off they can be added over the lines in red ink. This scale will then give accurate results if the capacity varies linearly with the setting, a condition which holds closely enough in the ordinary condensers. This same scale may also be affixed to the dial of any variable inductor and used without error if the variation of inductance with setting is linear. Also on either condenser or inductor, the same scale is used either with moving pointer and stationary dial or with moving dial.

A measurement of phase difference is made by first observing the current-square at resonance, then reading the scale at a setting on each side of resonance for which the current-square is one-half its value at resonance. The difference between the two readings on the scale is the value of ψ in degrees. The value of power factor in per cent. may be obtained from the result if desired by multiplying by 1.75.

A similar scale is readily made to read decrements directly. The readings of Figure 7 are all divided by 18.24, or the scale is independently calculated by equation (38). The scale shown in Figure 8 is thus obtained, which may be used on any condenser with semi-circular plates. Since writing this paper, the author has been informed that a scale constructed on this principle was devised for use in a decremeter by Mr. Waterman

of the Marconi Company. This is the only instance known in which the method has been used. It has here been shown to be convenient for measurements of other quantities than decrement, and is worthy of wide application, since it converts any wavemeter into a direct-reading phasemeter or decremeter at no additional cost.

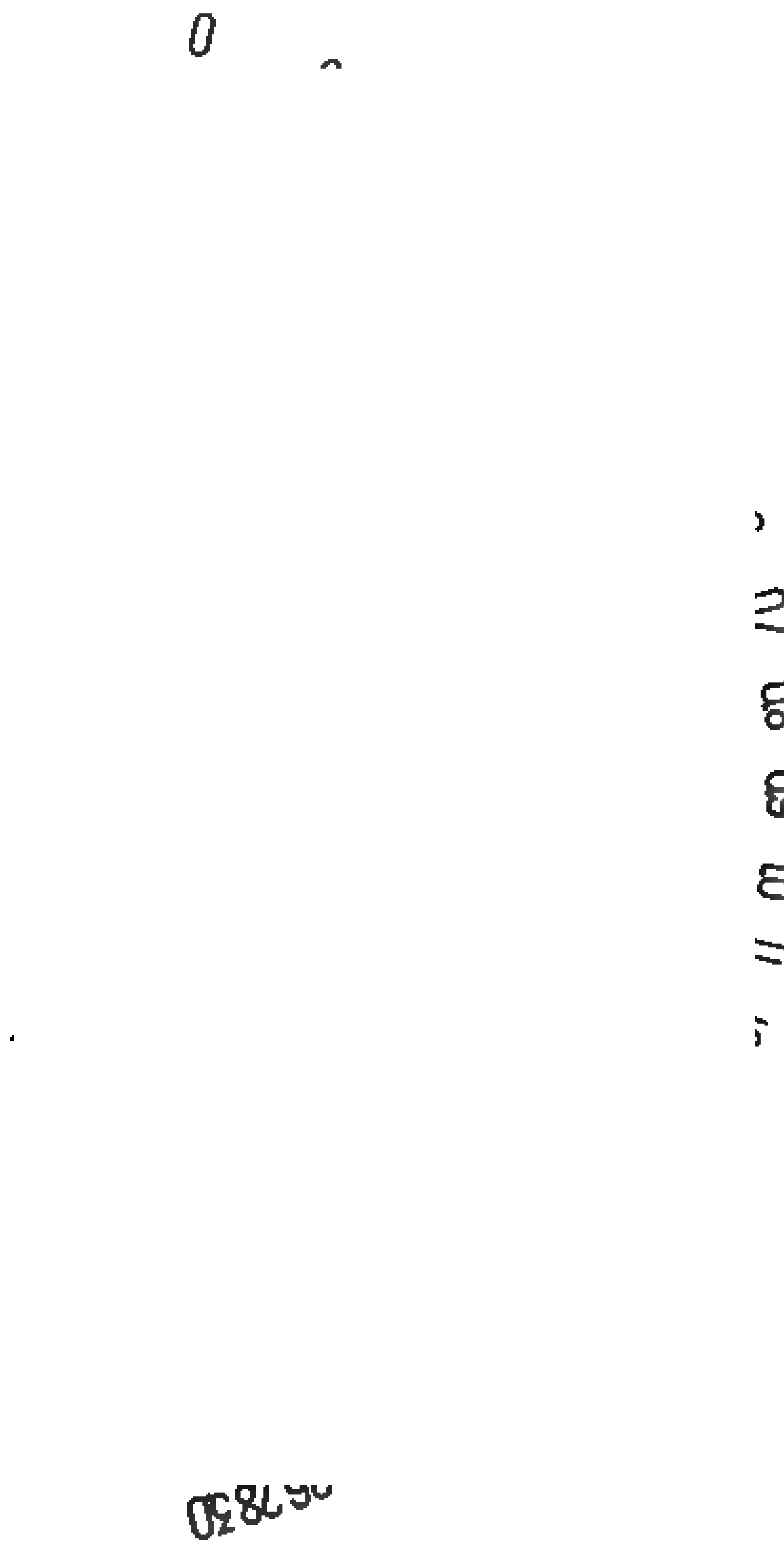

FIGURE 8

Location of Scale for Accurate Measurements

The scale gives accurate results only when $C_o=\beta_o=0$. In many semi-circular plate condensers C_o or β_o has a small positive or negative value. This can always be reduced to zero by shifting the β scale. Thus for a particular position of the scale suppose

$$\begin{aligned} C &= a\beta'+C_o \\ &= a(\beta'+\beta_o) \end{aligned} \tag{45}$$

Define a new β such that

$$\beta=\beta'+\beta_o \tag{46}$$

This reduces (45) to $C=a\,\beta$, and it is accomplished by shifting the dial toward zero by the amount β_o. The value of β_o is determined by two measurements of capacity. Suppose C_1 and C_2 are the values for β_1' and β_2'. The constant a is the change of capacity per degree and is given by

$$a=\frac{C_2-C_1}{\beta_2'-\beta_1'} \tag{47}$$

The angle β_o is therefore given by

$$\beta_o=\frac{C_1}{a}-\beta_1' \tag{48}$$

The scale of Figure 7 is accurately placed as follows. It is first placed on the dial by eye, and the capacity in the circuit accurately observed at the two points marked 5 and 36 on the scale. The amount by which the scale is to be shifted toward zero is then the angle in degrees,

$$\beta_o=\frac{100\,C_{36}}{C_5-C_{36}}-51.2 \tag{49}$$

The capacity concerned is the total capacity in the circuit, which consists mainly of the capacity of condenser and of the inductance coil in parallel with it. Since the coils of a wavemeter do not all have the same capacity, it is desirable to mount the phase difference scale in such a way that its angular position can be varied a few degrees on the dial, to correspond to the different coils that are used. A fiducial mark can be placed on the scale for each coil.

Measurement of Small Phase Difference or Decrement

These scales permit accurate measurement of fairly large phase differences or decrements, but offer no precision in the

measurement of very small values, particularly at the low-capacity end of the scale. They are thus of particular value in tests upon condensers or other apparatus having fairly large phase differences. The method can, however, be extended to the precise measurement of small values in several ways. One method is to use a gear to open out the scale. The scale can then be in the form of a spiral on the rapid-motion gear shaft, and be spaced by a factor equal to the gear ratio. This device does not have the simplicity of merely attaching a scale to the condenser dial. Another method is to place a condenser of fixed capacity in parallel with the variable and use a different scale on the variable. This narrows the range of capacity variation, but for some kinds of work the method is very satisfactory, and any desired precision of measurement may be obtained. The scale suitable for the use in parallel with the variable of a fixed capacity equal to 10 times that at the middle of the scale of the variable, is obtained as follows. The fixed capacity is C_o in (40), and its value must be equal to 10 times that at the 90° point on the variable condenser (or 19.86 on the scale of Figure 7). Expressing β in degrees, $C_o = a\ \beta_o = a \times 900$. Equation (41) becomes

$$s = \frac{2.303}{2}\left[\log_{10}(\beta_a + 900) - \log_{10}(\beta + 900)\right]$$

For the ψ scale beginning at the upper end of the capacity scale and s expressed in minutes, this becomes

$$s = \frac{60\,(57.3)\,(2.303)}{2}\left[\log_{10} 1080 - \log_{10}(\beta + 900)\right]$$

or,

$$\log_{10}(\beta + 900) = 3.033424 - \frac{s}{3957.9} \tag{50}$$

The scale thus calculated is given in Figure 9. It may be used on any linear scale condenser, as in the previous cases, and permits measurements of phase difference to closer than one minute. A similar scale is obtained for decrement by dividing the scale readings by 1094., permitting the measurement of decrement to better than 0.001. These scales have the additional advantage of almost uniform spacing.

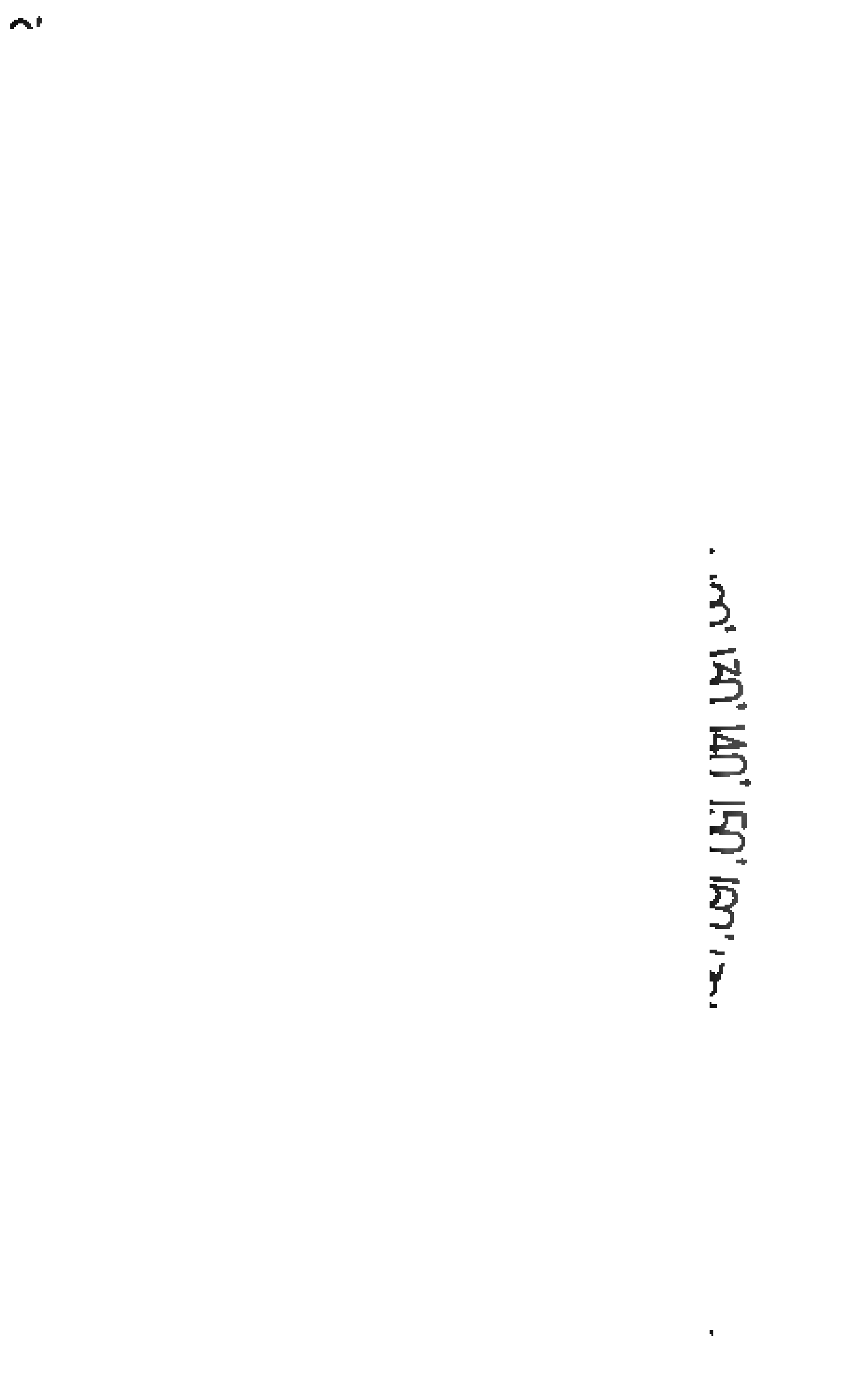

FIGURE 9

DECREMETER OR PHASEMETER WITH UNIFORM SCALE

Just as it is possible to determine a ψ or δ scale to fit a condenser having any sort of law of capacity variation, it is equally possible to design a condenser with capacity varying in such a way as to fit any specified ψ or δ scale. A uniform scale, i. e., one in which the graduations are equally spaced, is particularly convenient, and is the kind used in the Kolster decremeter.

A uniform scale of either ψ or δ requires in accordance with equation (34) that the condenser plates be so shaped that for any small variation of setting the ratio of the change in the capacity to the total capacity is constant. The condenser required to give this uniform scale has its moving plates so shaped that the logarithm of the capacity is proportional to the angle of rotation of the plates.

This decremeter is fully described in "Bulletin of the Bureau of Standards," **11**, page 421, 1914, Scientific Paper Number 235.* By the use of a separate shaft geared to the moving plates at a 6-to-1 ratio, the decrement scale is opened out so that very precise measurements may be made. This decremeter is used in the inspection service of the Bureau of Navigation of the Department of Commerce and by radio engineers elsewhere. On account of the uniform scale of decrements its use is more convenient than the instruments with specially shaped scales, but, on the other hand, the adjustment of the instrument to read decrements accurately is more difficult as this requires the adjustment of a small auxiliary condenser in parallel with the variable condenser. It is, of course, a more costly instrument because of the specially shaped condenser plates. It can be made to read phase difference directly in degrees by replacing the decrement scale with another in which the readings are multiplied by 18.24.

Use of Damped Oscillations

When damped oscillations are used in a measurement of resistance or one of the related quantities, there are two distinct decrements concerned, that of the circuit and that of the emf. supplied to the circuit. To determine either of these, in general two measurements are required. In special cases, however, one measurement only is necessary. For example, when the decrement of the supplied emf. is very small, the measurement of resistance of the circuit is made exactly the same as when the emf. is undamped and the equations are unchanged, in all the methods. Also, when impulse excitation is used for the resistance-variation method, the procedure is the same as with undamped emf.; the equations are different in this case, as shown in equation (16) above. In any of these cases, of course, the results of measurement can be expressed in terms of ψ, s, or δ of the circuit as well as R.

When the emf. supplied to the circuit has a moderate de-

*See also "Proc. Inst. Radio Engrs.," volume 3, number 1, page 29, 1915.

crement, damped oscillations flow in the circuit. Calculation of the current is very difficult except when the decrement of both the emf. and the circuit are small. The definition of decrement that has been given in terms of an energy ratio furnishes some interesting relations in this connection. Suppose the emf. is produced in the measuring circuit (Figure 5) by coupling to the source circuit so loosely that there is no reaction upon the source. If I' = the root-mean-square current of small decrement in the source circuit and M is the mutual inductance between the two circuits, it may be shown that the r.m.s value of emf. induced in the measuring circuit at resonance is

$$E = \omega M I'$$

and the maximum amplitude is

$$E_o = \omega M I_o'$$

From equation (9), $(I')^2 = \frac{N}{4f\delta'}(I_o')^2$,

therefore

$$E^2 = \frac{N}{4f\delta'} E_o^2$$

where δ' is the decrement of the current in the source circuit and hence of the emf. induced in the measuring circuit. Using $\frac{E^2}{R}$ as a definition of power consumption, the average power dissipated is

$$\frac{E^2}{R} = \frac{NE_o^2}{4f\delta' R}$$

Average energy dissipated per cycle $= \frac{NE_o^2}{4f^2\delta' R}$.

Average energy associated with current at maxima $= LI^2$, provided the decrement is small, as before. Assuming now that decrements are additive, and applying the energy-ratio definition given just after equation (9) to the sum of δ', the decrement of the applied emf. and δ, the decrement of the circuit,

$$\delta' + \delta = \frac{1}{2} \cdot \frac{NE_o^2}{4f^2 \delta' R L I^2}$$

$$I^2 = \frac{NE_o^2}{8f^2 R L \delta' (\delta' + \delta)}$$

$$I^2 = \frac{NE_o^2}{16 f^3 L^2 \delta' \delta (\delta' + \delta)} \qquad (51)$$

This is the correct relation between I^2 and E_o^2 at resonance,

as obtained from the elaborate rigorous proofs. The short demonstration just given involves the assumption that decrements are additive, which seems reasonable since energies are additive.

Resistance-Variation Method

Measurements made with damped waves are most conveniently expressed in terms of decrements. Resistances and the other related quantities can then be calculated from the values of decrement.

The equation for the resistance variation method is obtained from (51). Suppose the resistance of the circuit to be increased by an amount R_1 changing δ to $\delta+\delta_1$, and the original resonance current I to some other value I_1; then

$$I_1^2 = \frac{NE_o^2}{16 f^3 L^2 \delta' (\delta+\delta_1)(\delta'+\delta+\delta_1)} \cdot$$

$$\frac{I^2}{I_1^2} = \frac{(\delta+\delta_1)(\delta'+\delta+\delta_1)}{(\delta'+\delta)} \tag{52}$$

This is the equation for the resistance-variation method of measurement, using damped waves. It applies only when the decrements are small, and when the coupling to the source is so loose that the emf. is not affected by the current in the measuring circuit.

It is possible to solve either for δ' if δ is known or vice versa. When the method is thus used to obtain δ', the decrement of the applied emf., the result of the measurement really gives the shape of the trains of waves which are acting on the circuit. Decrement measurement may thus accomplish something similar at radio frequencies to what is done at low frequencies by wave analysis.

Determination of Decrement of Wave

The solution for δ' is

$$\delta' = \frac{2\,\delta\,\delta_1 + \delta_1^2 - \frac{I^2 - I_1^2}{I_1^2}\,\delta^2}{\frac{I^2 - I_1^2}{I_1^2}\,\delta - \delta_1} \tag{53}$$

This may be simplified by choosing the resistance inserted such that $\delta_1 = \delta$; then

$$\delta' = \delta\,\frac{4\,I_1^2 - I^2}{I^2 - 2\,I_1^2} \tag{54}$$

Another convenient simplified procedure is to vary the inserted resistance until the square of the current is reduced to one-half its previous value, then $\frac{I^2-I_1^2}{I_1^2}=1$, and

$$\delta'=\frac{2\,\delta\,\delta_1+\delta_1^2-\delta^2}{\delta-\delta_1} \tag{55}$$

This equation expresses the method presented by L. Cohen in "PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS," **2,** page 237, 1914.

DETERMINATION OF DECREMENT OF CIRCUIT

When δ' is the known quantity, the direct solution of (52) for δ, the decrement of the measuring circuit, is

$$\delta=\frac{1}{B}\delta_1-\frac{1}{2}\delta'\pm\frac{1}{2B}\sqrt{(B\,\delta')^2+4\,\delta_1^2+4B\,\delta_1^2} \tag{56}$$

where
$$B=\frac{I^2-I_1^2}{I_1^2}$$

This complicated form of solution is of very little use. Equation (52) is itself a more convenient expression than this explicit solution. The following formula has been found useful in certain cases as discussed below.

$$\delta=\delta_1\frac{K\,I_1^2}{I^2-K\,I_1^2} \tag{57}$$

where
$$K=1+\frac{\delta_1}{\delta'+\delta} \tag{58}$$

It is sometimes advantageous to express this in terms of resistance or the related quantities. Thus the solution for R of the circuit, where R_1 is the inserted resistance, is

$$R=R_1\frac{K\,I_1^2}{I^2-K\,I_1^2} \tag{59}$$

This is, of course, not an explicit solution for R, since K involves δ and, therefore, R, but gives a ready means for finding R or δ when the sum of the two decrements $(\delta'+\delta)$ is known from some other measurement, such as the reactance-variation method described below. Thus a combination of the two methods gives both δ' and δ, or δ' and R.

An interesting special case occurs when δ and δ_1 are both very small compared with δ'. K becomes unity and equation (59) reduces to

$$R=R_1\frac{I_1^2}{I^2-I_1^2} \tag{60}$$

This happens to be the same as equation (16) above, the equation for the use of impulse excitation. The proof given here can not, however, be regarded as a deduction of equation for impulse excitation, as it has been by some writers; since equation (51) is involved, which assumes that δ' and δ are both small.

Reactance-Variation Method

The procedure when the supplied emf. is damped is the same as when undamped, two observations of current being taken, one at resonance and the other after varying the reactance. The equations for decrement are only slightly different from those applying to undamped current.

Bjerknes' classical proof shows that the sum of the decrements of the emf. and of the measuring circuit is given by the same expression as that which gives the decrement of the measuring circuit when the emf. is undamped. Thus (28) becomes (61) below, and the equations for decrement corresponding to (21) to (25) become

$$\delta'+\delta=\pi\frac{\pm(C_r-C)}{C}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \qquad (61)$$

$$\delta'+\delta=\pi\frac{\pm(L-L_r)}{L_r}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \qquad (62)$$

$$\delta'+\delta=\pi\frac{\pm(\omega^2-\omega_r^2)}{\omega\,\omega_r}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \qquad (63)$$

$$\delta'+\delta=\pi\frac{\pm(\lambda_r^2-\lambda^2)}{\lambda_r\,\lambda}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \qquad (64)$$

$$\delta'+\delta=\pi\frac{C_2-C_1}{C_2+C_1}\sqrt{\frac{I_1^2}{I_r^2-I_1^2}} \qquad (65)$$

These formulas are correct only when: (1) the coupling between the source of emf. and measuring circuit is so loose that the latter does not appreciably affect the former; (2) δ' and δ are both small compared with 2π, and (3) the ratio $\frac{(C_r-C)}{C}$ and the corresponding ratios are small compared with unity. From any of these δ' is obtained if δ is known and vice versa. If a separate measurement is made by the method of equation (57) above, both decrements are obtained.

The appearance of the sum $(\delta+'\delta)$ in the equations does not mean that the current flowing in the measuring circuit actually has a decrement equal to $(\delta'+\delta)$. As a matter of fact the actual decrement of the current is a value nearly equal to which-

ever of the two, δ' or δ, is the smaller. For this reason the equations involving $(\delta'+\delta)$ can not be extended to the measurement of the sum of the decrements of two loosely coupled circuits by coupling to one of them a third measuring circuit, as has sometimes been tried.

As mentioned earlier, the reactance-variation method is simplified if the reactance is varied by such an amount as to make $I_1^2 = \frac{1}{2} I_r^2$. This is done very easily when the current measuring instrument is graduated in terms of current squared. The quantity under the square root sign in all the preceding equations becomes unity, greatly simplifying the formulas. Calculation may be entirely eliminated by use of direct-reading decremeters as previously described. Such instruments when thus used with damped waves give directly $(\delta'+\delta)$.

SUMMARY: The methods of measuring resistance and related quantities at radio frequencies are fewer in number and necessarily different from those at low frequencies. The conditions of such measurements and the relations of various methods have not previously been given in comprehensive fashion. This paper shows the relations between resistance, phase difference, sharpness of resonance, and decrement. The methods of measurement are derived and classified. Most of the valuable methods are comprised under the resistance-variation and reactance-variation methods. Special direct-reading methods of measuring phase difference and decrement are presented.*

*Extra copies of the special scales of figures 7, 8, 9 can be obtained from The Institute of Radio Engineers by addressing the Editor, The College of the City of New York.

SYMBOLS USED IN THIS PAPER

C = capacity of condenser in measuring circuit.
C' = capacity of condenser in source circuit.
d = deflection of current measuring instrument.
d_1 = deflection when known resistance is inserted in circuit.
E = effective electromotive force.
E_o = maximum electromotive force.
f = frequency of alternation.
I = effective current.
I_o = maximum current.
I_r = current at resonance.
I_1 = current when either resistance or reactance of circuit is increased.
L = self-inductance of circuit.
M = mutual inductance.
N = number of trains of oscillations per second.
P = average power.
r = subscript used to denote resonance.
R = resistance of circuit.
R_1 = known resistance inserted in circuit.
R_x = resistance of apparatus under measurement.
s = scale setting.
S = sharpness of resonance
X = reactance.
X_1 = change of reactance.
W = average energy.
α = damping factor.
δ = logarithmic decrement of circuit.
δ_1 = increase in decrement caused by adding resistance.
δ' = decrement of applied emf.
ε = base of napierian logarithms = 2.71828.
θ = phase angle of C or of L, considering the resistance to be associated with it.
λ = wave length.
ψ = phase difference of C or L, considering the resistance to be associated with it.
$\omega = 2\pi \times$ frequency.

NOTE ON LOSSES IN SHEET IRON AT RADIO FREQUENCIES*

By

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(Paris, France)

I. The study of Foucault currents in iron sheets at high frequency was made first by Oliver Heaviside, and later by J. J. Thomson.[1]

Recently, Mr. Bethenod[2] has considered the complication caused by the phenomenon of hysteresis, and has introduced that phenomenon in his calculations by utilizing the method of procedure first employed by Ferraris (1888), according to which hysteresis is supposed to cause a constant lag τ of phase-angle between the magnetic induction $\boldsymbol{B}$ and the magnetizing field $\boldsymbol{H}$. In seeking to determine the final phase-lag between the emf. and the current in a coil having a closed magnetic circuit, when the frequency is increased indefinitely, Mr. Bethenod has shown that this limiting phase-lag, instead of being equal to $\frac{\pi}{4}$, as indicated by the formulas of J. J. Thomson, is diminished by an angle equal to $\frac{\tau}{2}$ by the effect of hysteresis. In reality, Mr. Bethenod's conclusion, according to which the limiting angle of phase-lag, between the emf. and the current, is diminished thru hysteresis by an angle equal to $\frac{\tau}{2}$, does not seem to be related to any particular interpretation of the phenomenon of hysteresis. It is natural, in fact, that the losses due to hysteresis, as with the losses due to Foucault currents, should tend to increase the "watt" current absorbed, and, consequently, should tend to bring the current more nearly in phase with the emf.

Before proceeding to any calculation, it is easy to under-

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[1] "The Electrician," volume 28, 1892, page 599.

[2] "La Lumière Electrique," July 22, 1916, volume 34, 2nd Series, page 73.

stand how hysteresis influences Foucault currents, and to understand how Foucault currents may influence hysteresis losses. If we assume that hysteresis introduces a phase-lag between magnetic induction and ampere-turns, we can understand at once that the introduction of Foucault currents into the equations is thereby affected. On the other hand, the presence of Foucault currents causes the magnetic induction in iron sheets to vary from the center to the external surface; and since the losses due to hysteresis increase according to a power of the magnetic induction which is higher than the first power, the losses will be higher than if the magnetic induction in the sheet were assumed to remain uniform. The hysteresis losses, therefore, depend on the influence of Foucault currents on the distribution of magnetic induction.

It will be principally necessary to take into consideration this distribution of magnetic induction in the sheet in order to calculate the losses due to hysteresis. In particular, if we assume these losses to be proportional to the square of the magnetic induction, which is all the more likely because, at high frequency, the magnetic induction is always low, it will be necessary to know, somehow, the effective spatial distribution of the magnetic induction in the sheet.

The purpose of the author is to revise the methods of dealing mathematically with Foucault currents, and to study the influence of hysteresis on losses due to Foucault currents, as well as to study the losses due to hysteresis itself. His purpose also is to obtain formulas which are useful for the study of radio frequency apparatus.

II. To establish the equations for Foucault currents, the author will not introduce directly the general equations of Maxwell, as is usually done; these equations can be established by equivalent considerations which are much more familiar to engineers.

Let us consider (Figure 1) a sheet of thickness $2a$. Let us take as origin the median plane XX, parallel to the two faces of the sheet, S and S'. On account of symmetry, the Foucault currents which go thru the sheet will give rise to currents of equal densities, and of opposite polarities at symmetrical points x and x', situated at equal distances, Ox and Ox', from the origin O. The Foucault currents which tend to form a shield against the magnetic flux which is passing thru the sheet, in a direction parallel to the faces S and S', follow the directions indicated

by arrows in the plane of the figure. It should be noted, moreover, that the ampere-turns acting to produce a magnetic field at the point x, are those due to currents circulating in the region at the right of Ox and at the left of Ox'. The currents circulating

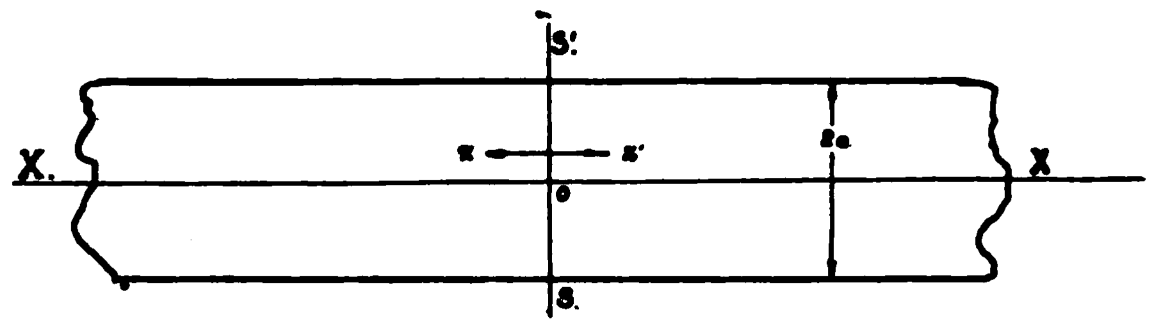

FIGURE 1

within the space bounded by the planes parallel to XX, which pass thru x and x', produce, in fact, no field external to that space. Let us designate by δ the current-density in the sheet at the point x. The decrease in ampere-turns per centimeter, which results in passing from a thickness dx toward the external surface S of the sheet, is $\delta\, dx$. The magnetic induction $\boldsymbol{B}$ at the point x will be decreased by a corresponding amount $d\boldsymbol{B}$ such that

$$d\boldsymbol{B} = -4\pi\mu\delta\, dx$$

whence

$$\frac{d\boldsymbol{B}}{dx} = -4\pi\mu\delta \tag{1}$$

In reality, the current density δ and the magnetic induction B, being both harmonic functions of time, can be expressed as follows:

$$\delta = \delta_1 \sin \omega t - \delta_2 \cos \omega t$$

$$\boldsymbol{B} = \boldsymbol{B}_1 \sin \omega t - \boldsymbol{B}_2 \cos \omega t$$

in which equations any origin can be taken arbitrarily for time values (t). Under those conditions equation (1) corresponds to the two following equations:

$$\left.\begin{aligned} \frac{d\boldsymbol{B}_1}{dx} &= -4\pi\mu\delta_1 \\ \frac{d\boldsymbol{B}_2}{dx} &= -4\pi\mu\delta_2 \end{aligned}\right\} \tag{2}$$

If we assume that hysteresis introduces an angle of phase-lag, τ, between magnetic induction and ampere-turns, we can

say that everything happens, as far as the ampere-turns are concerned, as if the current density, δ, corresponded to a fictitious density, δ, having a phase-lag, τ, such that

$$\delta' = \delta_1 \sin(\omega t - \tau) - \delta_2 \cos(\omega t - \tau)$$
$$= (\delta_1 \cos\tau - \delta_2 \sin\tau) \sin\omega t - (\delta_2 \cos\tau + \delta_1 \sin\tau) \cos\omega t$$

Equations (2) then become

$$\begin{aligned} \frac{d\boldsymbol{B}_1}{dx} &= -4\pi\mu(\delta_1 \cos\tau - \delta_2 \sin\tau) \\ \frac{d\boldsymbol{B}_2}{dx} &= -4\pi\mu(\delta_2 \cos\tau + \delta_1 \sin\tau) \end{aligned} \tag{2'}$$

The choice between equations (2) and (2′) will then depend on whether the phenomenon of hysteresis is to be taken into consideration or not, in the equations for Foucault currents.

We now proceed to establish a second equation by a simple consideration which follows.

The emf. induced per centimeter in the direction XX between the two planes which pass thru the abscissa points x and $x+dx$ is

$$-\frac{d}{dt}\boldsymbol{B}\,dx = -\omega(\boldsymbol{B}_1 \cos\omega t + \boldsymbol{B}_2 \sin\omega t)\,dx$$

This emf. must be exactly balanced by the difference between the ohmic drop per centimeter in the plane passing thru x and the ohmic drop per centimeter in the plane passing thru $x+dx$. This difference is equal to $\varrho\, d\,\delta$, where δ designates the resistivity of the sheet. We therefore have

$$\begin{aligned} -\omega(\boldsymbol{B}_1 \cos\omega t + \boldsymbol{B}_2 \sin\omega t)\,dx &= \varrho\, d\,\delta \\ &= \varrho\left(\frac{d\delta_1}{dx} \cdot \sin\omega t - \frac{d\delta_2}{dx}\cos\omega t\right)dx \end{aligned} \tag{3}$$

From this we obtain the two following equations:

$$\begin{aligned} \omega\boldsymbol{B}_1 &= \varrho\frac{d\delta_2}{dx} \\ \omega\boldsymbol{B}_2 &= -\varrho\frac{d\delta_1}{dx} \end{aligned} \tag{4}$$

From equations (2′) and (4) we obtain the two following equations of the second order:

$$\begin{aligned} \frac{d^2\delta_1}{dx^2} &= \frac{4\pi\mu\omega}{\varrho}(\delta_2 \cos\tau + \delta_1 \sin\tau) \\ \frac{d^2\delta_2}{dx^2} &= -\frac{4\pi\mu\omega}{\varrho}(\delta_1 \cos\tau - \delta_2 \sin\tau) \end{aligned} \tag{5}$$

Taking

$$\frac{4\pi\mu\omega}{\varrho}=2m^2,\quad \sqrt{1+\sin\tau}=\alpha,\quad \sqrt{1-\sin\tau}=\beta$$

we obtain, by integration, the following:

$$\begin{aligned}\delta_1&=-A\,\frac{e^{m\alpha x}-e^{-m\alpha x}}{2}\cos m\beta x=-A\sinh m\alpha x\cos m\beta x\\ \delta_2&=A\,\frac{e^{m\alpha x}+e^{-m\alpha x}}{2}\sin m\beta x=A\cosh m\alpha x\sin m\beta x\end{aligned}\tag{6}$$

in which A is a constant of integration.[3]

From these and from equation (4), the values of $\boldsymbol{B}_1$ and $\mathbf{B}_2$ are obtained:

$$\begin{aligned}\boldsymbol{B}_1&=\frac{A\varrho m}{\omega}(\alpha\sinh m\alpha x\sin m\beta x+\beta\cosh m\alpha x\cos m\beta x)\\ \boldsymbol{B}_2&=-\frac{A\varrho m}{\omega}(\beta\sinh m\alpha x\sin m\beta x-\alpha\cosh m\alpha x\cos m\beta x)\end{aligned}\tag{7}$$

From the above solution (6) we obtain the maximum current density value δ_{max} at the point x:

$$\delta_{max}=\sqrt{\delta_1^2+\delta_2^2}=\frac{A}{\sqrt{2}}\sqrt{\cosh 2m\alpha x+\cos 2m\beta x}\tag{8}$$

From solution (7) we obtain the maximum value of magnetic induction $\boldsymbol{B}_{max}$ at the point x:

$$\boldsymbol{B}_{max}=\frac{A\varrho m}{\omega}\sqrt{\cosh 2m\alpha x+\cos 2m\beta x}\tag{9}$$

We can now determine the constant A by starting either from the apparent, or the mean magnetic induction in the sheet, or from the external ampere-turns per centimeter (J) which act on the sheet.

When starting from the apparent magnetic induction $\boldsymbol{B}_{app}$ it is to be noted that the emf. per centimeter, which must equal the ohmic drop due to Foucault currents along the external surface of the sheet, is $\omega\boldsymbol{B}_{app}a$. We therefore have:

$$\omega\boldsymbol{B}_{app}a=\frac{\varrho A}{\sqrt{2}}(\cosh 2m\alpha a-\cos 2m\beta a)^{\frac{1}{2}}$$

Whence:

$$A=\frac{\sqrt{2}\,\omega a\boldsymbol{B}_{app}}{\varrho(\cosh 2m\alpha a-\cos 2m\beta a)^{\frac{1}{2}}}\tag{10}$$

[3] The general integral would require a second constant, but it is found that this constant must be equal to zero in order that the condition of symmetry implying $\delta_1=\delta_2=0$ should be satisfied for $x=0$.

When starting from the external ampere-turns, J, per centimeter, it must be noted that the magnetic induction at the surface of the sheet must be equal to $4\pi\mu J$. We therefore have:

$$4\pi\mu J = \frac{A\varrho m}{\omega}(\cosh 2m\alpha a + \cos 2m\beta a)^{\frac{1}{2}}$$

Whence:

$$A = \frac{\omega}{\varrho m}\frac{4\pi\mu J}{(\cosh 2m\alpha a + \cos 2m\beta a)^{\frac{1}{2}}} \tag{11}$$

Apparent Permeability.—Equating (10) and (11) we have:

$$\frac{\sqrt{2}\,a\,B_{app}}{(\cosh 2m\alpha a - \cos 2m\beta a)} = \frac{4\pi\mu J}{m(\cosh 2m\alpha a + \cos 2m\beta a)^{\frac{1}{2}}}$$

From this we obtain immediately an expression for the apparent permeability:

$$\mu_{app} = \frac{\mu}{\sqrt{2}\,m\,a}\frac{(\cosh 2m\alpha a - \cos 2m\beta a)^{\frac{1}{2}}}{(\cosh 2m\alpha a + \cos 2m\beta a)^{\frac{1}{2}}} \tag{12}$$

This expression for the apparent permeability becomes identical with that given by J. J. Thomson when we assume $\tau = 0$, that is to say $\alpha = \beta = 1$.

It is necessary to know the expression for the apparent permeability, in order to determine the given or apparent magnetic induction, as a function of the available ampere-turns, in any given radio frequency apparatus.

III. We now proceed to determine the Foucault current and hysteresis losses as a function of the mean or apparent magnetic induction.

Foucault Current Losses.—In order to determine the Foucault current losses, it is necessary to determine in some manner the effective value in space of the current density δ_{max}, that is to say, the value of:

$$\frac{1}{a}\int_0^a \delta_{max}^2\,dx$$

We have

$$(\delta_{max}^2)_{av} = \frac{A^2}{2a}\int_0^a(\cosh 2m\alpha x - \cos 2m\beta x)\,dx$$

$$= \frac{A^2}{4ma}\left(\frac{\sinh 2m\alpha a}{\alpha} - \frac{\sin 2m\beta a}{\beta}\right)$$

Bearing in mind that the effective current density, as a func-

tion of the time, is $\sqrt{2}$ times lower than the maximum density, the Foucault current losses per cubic centimeter will be:

$$W_F = \frac{A^2 \varrho}{8\, m\, a}\left(\frac{\sinh 2\, m\, a\, a}{a} - \frac{\sin 2\, m\, \beta\, a}{\beta}\right) .$$

If we replace the constant A by its value obtained from (10), we have,

$$W_F = \frac{\omega^2\, a}{4\, m\, \varrho}\, \frac{\dfrac{\sinh 2maa}{a} - \dfrac{\sin 2\, m\, \beta\, a}{\beta}}{\cosh 2\, m\, a a - \cos 2\, m\, \beta\, a}\, B^2_{app} \tag{13}$$

The effect of hysteresis on Foucault current losses is readily seen. Leaving out hysteresis, the expression of these losses takes the simple form:

$$W_F = \frac{\omega^2\, a}{4 m \varrho}\, \frac{\sinh 2\, m\, a - \sin 2\, m\, a}{\cosh 2\, m\, a - \cos 2\, m\, a}\, B^2_{app} . \tag{13'}$$

Hysteresis Losses.—To determine the hysteresis losses, we must bear in mind that the assumption of a constant phase-lag τ of the magnetic induction $\boldsymbol{B}$ behind the magnetizing field $\boldsymbol{H}$ supposes the hysteresis losses to be proportional to the square of the maximum induction. It is therefore necessary to determine in some way the effective value of $\boldsymbol{B}_{max}$ in space.

We have:

$$(\boldsymbol{B}^2_{max})_{av} = \frac{A^2\, \varrho^2\, m^2}{\omega^2\, a} \int_0^a (\cosh 2\, m a x + \cos 2\, m \beta x)\, d\, x$$

$$= \frac{A^2\, \varrho^2\, m^2}{2\, \omega^2\, a}\left(\frac{\sinh 2\, m\, a\, a}{a} + \frac{\sin 2\, m\, \beta\, a}{\beta}\right) .$$

We also know that the same hypothesis of constant phase-angle between the magnetic induction $\boldsymbol{B}$ and the magnetizing field $\boldsymbol{H}$ implies that, in the formula

$$W = \eta\, \boldsymbol{B}^2_{max}$$

which gives the hysteresis losses per cycle, the value of the coefficient η is equal to $\dfrac{\sin \tau}{4\, \mu}$. Under those conditions, the hysteresis losses per cubic centimeter are equal to:

$$W_H = \frac{\sin \tau}{4\, \mu}\, \frac{\omega}{2\, \pi}\, \frac{A^2\, \varrho^2\, m}{2\, \omega^2\, a}\left(\frac{\sinh 2\, m\, a\, a}{a} + \frac{\sin 2\, m\, \beta\, a}{\beta}\right)$$

$$= \frac{\sin \tau}{8}\, \frac{A^2 \varrho}{m a}\left(\frac{\sinh 2\, m\, a\, a}{a} + \frac{\sin 2\, m\, \beta\, a}{\beta}\right)$$

If we replace the constant A by its value from (10), we have:

$$W_H = \frac{\omega^2 a \sin\tau}{4 m \varrho} \frac{\dfrac{\sinh 2 m \alpha a}{\alpha} + \dfrac{\sin 2 m \beta a}{\beta}}{\cosh 2 m \alpha a - \cos 2 m \beta a} B^2_{app} \tag{14}$$

Ratio of Losses.—The ratio of hysteresis losses to Foucault current losses can be readily determined. We find:

$$\frac{W_H}{W_F} = \frac{\dfrac{\sinh 2 m \alpha a}{\alpha} + \dfrac{\sin 2 m \beta a}{\beta}}{\dfrac{\sinh 2 m \alpha a}{\alpha} - \dfrac{\sin 2 m \beta a}{\beta}} \sin\tau \tag{15}$$

This ratio tends toward *versin* τ in proportion as the frequency increases.

Total Losses.—The expression for the total losses takes the form:

$$W_F + W_H = \frac{\omega^2 a}{4 m \varrho} \frac{\alpha \sinh 2 m \alpha a - \beta \sin \alpha m \beta a}{\cosh 2 m \alpha a - \cos 2 m \beta a} B^2_{app} \tag{16}$$

If the calculation were not complicated by introducing the phase-angle τ, and by seeking to determine the hysteresis losses according to the formula for losses per cycle (10), equation (14) could be simplified by replacing $\sin\tau$ by its value $4\mu\eta$, and by making $\tau = 0$, that is to say, by assuming $\alpha = \beta = 1$ everywhere else in the equation; which would give the following value for W_H,

$$W_H = \frac{\mu \eta \omega^2 a}{\varrho m} \frac{\sinh 2 m a + \sin 2 m a}{\cosh 2 m a - \cos 2 m a} B^2_{app}. \tag{14'}$$

Taking into consideration the uncertainty which exists in regard to the exact value of hysteresis losses per cycle, and owing to the circumstances that all authors differ as to the value which should be given to the exponent of B, to which the losses are proportional, the simplified expression for W_H, given in equation (14′), may often be utilized.

Under those conditions, the total losses derived from (13′), and (14′), will be as follows:

$$W_F + W_H = \frac{\omega^2 a (1 + 4 \eta \mu) \sinh 2 m a}{4 m \varrho \cosh 2 m a - \cos 2 m a} B^2_{app}. \tag{16'}$$

Minimum Losses.—It is possible to determine the losses per cubic centimeter in a total volume containing the iron sheets and the insulation between them.

Let ε be the thickness of the insulation between the sheets,

and B'_{app} the mean magnetic induction in the total section composed of the sheets and of the insulation between them. The apparent induction in the sheet itself will be:

$$\frac{2a+\varepsilon}{2a} B'_{app}$$

On the other hand, the space occupied by the iron will be reduced in the proportion $\frac{2a}{2a+\varepsilon}$. Finally, the losses per cubic centimeter of total space will be, from (16), as follows:

$$W_F + W_H = \frac{\omega^2 (2a+\varepsilon)}{8m\varrho} \frac{\alpha \sinh 2m\alpha a - \beta \sin 2m\beta a}{\cosh 2m\alpha a - \cos 2m\beta a} B'^2_{app} \qquad (17)$$

The thickness of sheets $2\,a_{opt}$ which will give the minimum of loss in a given volume for a given magnetic induction B'_{app} will be that which will always give the minimum value to the preceding expression. This minimum value is obtained by taking the derivative of that expression with respect to "a." In other words, it will be that thickness which represents the solution of the following transcendental equation:

$$2m(2a+\varepsilon)\frac{1-\cosh 2m\alpha a \cos 2m\beta a}{\cosh 2m\alpha a - \cos 2m\beta a} + \alpha \sinh 2m\alpha a - \beta \sin 2m\beta a = 0 \qquad (18)$$

As a numerical illustration, let us take:

$$\mu = 2000$$
$$\varrho = 4\,(10)^4$$
$$\omega = 2\pi \times 30{,}000$$
$$\varepsilon = 0.003$$

From this we have:

$$m = \sqrt{\frac{2\pi\mu\omega}{\varrho}} = 243.5$$

We then find for $2\,a_{opt}$ the following values:

With $\sin\tau = 0.2$ $\quad 2\,a_{opt} = 0.027$ mm. (0.001 inch)

With $\sin\tau = 0.3$ $\quad 2\,a_{opt} = 0.0315$ mm. (0.0012 inch)

With $\sin\tau = 0.5$ $\quad 2\,a_{opt} = 0.0375$ mm. (0.0014 inch)

Phase-Angle Between EMF. and Current in an Inductance Having a Closed Magnetic Circuit

The losses have been evaluated directly without seeking to determine, as is usually done, the phase-angle ϕ between the emf. and the current. When the losses are known this

phase-angle can be determined quite easily by an inverse process.

The maximum emf. "V," induced per turn per square centimeter will be $\omega \boldsymbol{B}_{app}$, and the power consumed per cubic centimeter will, therefore, be:

$$\frac{VJ}{2} \cos \phi = \frac{\omega \boldsymbol{B}_{app} J}{2} \cos \phi$$

On the other hand, we know this power from equation (16) and we also know the value of $\boldsymbol{B}_{app}$, as a function of J, from that of the apparent permeability. Equating these two expressions for losses, the value of $\cos \phi$ may be readily obtained. We will have:

$$\cos \phi = \frac{1}{\sqrt{2}} \frac{a \sinh 2 m a a - \beta \sin 2 m \beta a}{(\cosh^2 2 m a a - \cos^2 2 m \beta a^{\frac{1}{2}}}$$

When m tends toward infinity $\cos \phi$ tends toward $\frac{a}{\sqrt{2}}$; or $\sqrt{\frac{1+\sin \tau}{2}}$; that is to say, ϕ tends toward the angle $\frac{\pi}{4} - \frac{\tau}{2}$, as Mr. Bethenod has shown. We have in fact:

$$\sqrt{\frac{1+\sin \tau}{2}} = \cos \left(\frac{\pi}{4} - \frac{\tau}{2}\right)$$

Phase-Angle Between EMF. and Current in a Coil with Open Magnetic Circuit

In a coil having an open magnetic circuit, which includes an air-gap that multiplies the apparent reluctance of the magnetic circuit by k, the magnetic induction for the same ampere-turns, J, will be divided by k. Consequently, the losses are divided by k^2, while the emf. induced at the terminals is divided by k. Under those conditions it will be found that the phase-angle ϕ' becomes such that:

$$\cos \phi' = \frac{\cos \phi}{k}$$

in which $\cos \phi$ retains the value indicated in the preceding paragraph.

In proportion as the air-gap is increased, the emf. and the current tend more and more to assume the quarter-phase relation. As a rule, it is $\tan \phi'$ which should be given as high a value as possible.

The author will return later to the important question of the construction of inductance coils with low losses (with or without iron) for a given frequency.

SUMMARY: In this article there is determined the power dissipated separately by Foucault currents and by hysteresis, in a sheet of iron, on the assumption that there exists a constant angle of lag between the magnetic induction in the sheet and the magnetizing field producing it. There is deduced the thickness which should be given to the iron sheets of apparatus supplied with radio frequency current, in order that the total power expended shall be a minimum. A calculation is made of the angle of lag between the voltage and the current in the circuit of an inductance coil.

THE NATURAL FREQUENCY OF AN ELECTRIC CIRCUIT HAVING AN IRON MAGNETIC CIRCUIT*

By

H. G. Cordes

(Research Engineer, Bremerton, Washington)

The natural frequency of a circuit depends upon inductance and capacitance and is modified by resistance and conductance in the circuit. When the inductance consists of a coil which has an iron magnetic circuit, the effective value of the inductance of the coil becomes a variable. The change in the value of the inductance is attributed to eddy currents induced in the iron. The eddy currents vary with the conductivity and permeability of the iron and the thickness of the laminations.

Other factors remaining constant, the frequency of an oscillatory circuit depends upon the value of the inductance and the effective inductance depends upon the frequency of the oscillating current in the circuit. The problem to be solved consists in finding an expression showing a relation between these two interdependent quantities.

The application of such an expression is illustrated by the following figure.

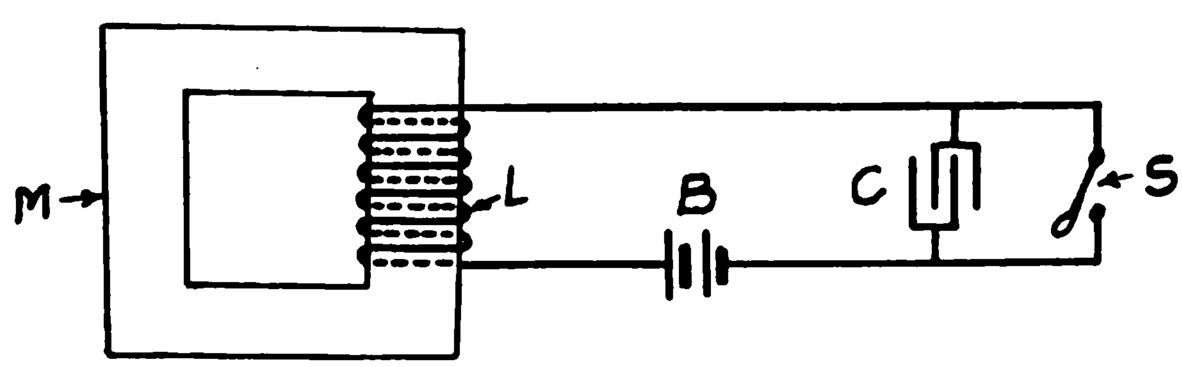

Figure 1

Magnetic flux is induced in the iron magnetic circuit M when current flows thru the coil L. Let B represent a battery, C a condenser and S a switch. When S is closed current passes thru

* Received by the Editor, April 16, 1918.

B, L, and S. Let r = the resistance of the circuit, and assume the potential of C to be zero.

Open the switch S quickly so that no sparking takes place. In order to determine the rate at which the potential rises in C and the maximum potential to which C will become charged, it is necessary to know the inductance of L, the capacitance of C, the resistance r and the initial current I flowing in the circuit.

Consider the potential of B and the potential drop $I\,r$ to be negligible compared with the emf. of self-induction. The electromagnetic energy initially stored in the coil is $\frac{1}{2} L_1 I^2$, where I is the initial current and L_1 is the inductance (coefficient of self-induction) of the coil at a low rate of change of the current. The value of L_1 decreases to an effective value L_m, as the average rate of change of current is increased due to an increase of eddy currents induced in the iron. Part of the initial electromagnetic energy is transformed into electrostatic energy, and the remainder is dissipated in the form of heat by eddy currents. The effective inductance of the coil is directly proportional to the total flux induced by the current thru the coil.

A method developed by Dr. C. P. Steinmetz for determining the ratio of the magnetic flux density due to a continuous current, to the effective magnetic flux density at a given frequency of alternating magnetic flux, is as follows:

Figure 2 shows the section of three laminations the thickness of which $= 2\,l_o$ and length $= S$. The thickness $2\,l_o$ is negligible compared to the length S. Current flows thru the conductor Z and in the direction of the arrow at the instant considered. The dotted lines show the path along which magnetic flux is induced

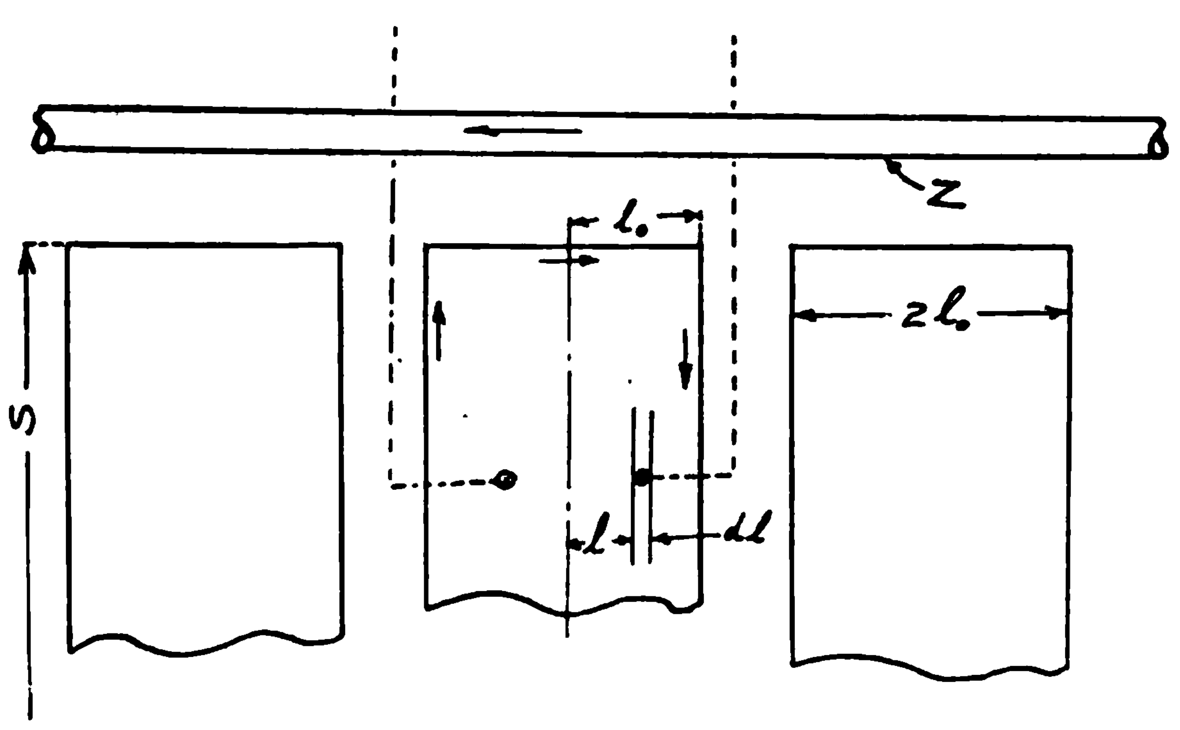

FIGURE 2

by the current. The flux induced in the iron is large compared with that induced in the air between the surface of the lamination and the surface of the conductor. The latter will be neglected and the total flux induced will be considered to generate an emf. in the iron at the surface of the lamination.

To obtain an expression for the flux density which is induced by the resultant magnetomotive force of the current in the conductor and the eddy current in the iron the instantaneous direction of which is shown by arrows, the following is quoted from Steinmetz's "Transient Electric Phenomena and Oscillations," Chapter VI (first edition):

"Let μ = the magnetic permeability, λ = the electric conductivity, l = the distance of a layer $d\,l$ from the center line of the lamination, and $2\,l_o$ = the total thickness of the lamination. If then I = the current density in the layer $d\,l$ and E = the emf. per unit length generated in the zone $d\,l$ by the alternating magnetic flux, we have

$$I = \lambda E \tag{1}$$

The magnetic flux density B_1 at the surface $l = l_o$ of the lamination corresponds to the impressed or external mmf. The density B in the zone $d\,l$ corresponds to the impressed mmf. plus the sum of all the mmf.'s in the zones outside of $d\,l$ or from l to l_o.

The current in the zone $d\,l$ is

$$I\,d\,l = \lambda E\,d\,l \tag{2}$$

and produces the mmf.

$$H = 0.4\,\pi\,\lambda E\,d\,l \tag{3}$$

which in turn would produce the magnetic flux density

$$dB = 0.4\,\pi\,\lambda\,\mu E\,d\,l \tag{4}$$

that is, the magnetic flux density B at the two sides of the zone $d\,l$ differs by the magnetic flux density dB (equation (4) produced by the mmf. in zone $d\,l$, and this gives the differential equation between B, E and l,

$$\frac{dB}{d\,l} = 0.4\,\pi\,\lambda\,\mu E \tag{5}$$

The emf. generated at distance l from the center of the lamination is due to the magnetic flux in the space from l to l_o. Thus the emfs. at the two sides of the zone $d\,l$ differ from each other by the emf. generated by the magnetic flux $B\,d\,l$ in this zone.

Considering now $\dot{B}, \dot{E}$ and $\dot{I}$ as complex quantities, the emf. $d\dot{E}$, that is, the difference between the emf.'s at the two sides of the zone $d\,l$, is in quadrature ahead of $\dot{B}\,d\,l$, and thus denoted by

$$d\dot{E} = -j\,2\,\pi f \dot{B}\,10^{-8}\,d\,l \tag{6}$$

where f is the frequency of the alternating magnetism.

This gives the second differential equation,

$$\frac{d\dot{E}}{d\,l} = -j\,2\,\pi f \dot{B}\,10^{-8} \tag{7}$$

The reader is referred to the text quoted for the steps by which the expression for the average flux density in the iron is obtained. It is briefly as follows:

From (5) and (6)

$$\dot{B} = \frac{\dot{B}_1}{2}\,\frac{\varepsilon^{(1-j)c\,l} + \varepsilon^{-(1-j)c\,l}}{\varepsilon^{(1-j)c\,l_o} + \varepsilon^{-(1-j)c\,l_o}} \tag{8}$$

where

$$c = \sqrt{0.4\,\pi^2 f\,\lambda\,\mu\,10^{-8}} \tag{9}$$

The average value of the flux density in the iron is

$$\dot{B}_m = \frac{1}{l_o}\int_o^{l_o} \dot{B}\,d\,l \tag{10}$$

Equation (8) in (10) gives

$$\dot{B}_m = \frac{\dot{B}_1}{(1-j)\,c\,l_o}\cdot\frac{\varepsilon^{(1-j)c\,l_o} - \varepsilon^{-(1-j)c\,l_o}}{\varepsilon^{(1-j)c\,l_o} + \varepsilon^{-(1-j)c\,l_o}} \tag{11}$$

The absolute value of $\dot{B}_m$ is the square root of the sum of the squares of the real and imaginary terms in equation (11), which, substituting hyperbolic and circular functions, is

$$B_m = \frac{B_1}{c\,l_o\sqrt{2}}\cdot\sqrt{\frac{\cosh 2\,c\,l_o - \cos 2\,c\,l_o}{\cosh 2\,c\,l_o + \cos 2\,c\,l_o}} \tag{12}$$

The inductance of a conductor or coil is directly proportional to the total magnetic flux induced by the current in the conductor or coil, therefore

$$\frac{L_m}{L_1} = \frac{B_m}{B_1} \tag{13}$$

The inductance L_1 is due to the flux induced in the iron by the mmf. of the current in the conductor or coil, while the inductance L_m is due to the flux induced in the iron by the resultant mmf.'s of the current in the conductor or coil and the eddy currents in the iron.

When the conductor Z is placed at an appreciable distance from the iron or the distance between the laminations is not negligible compared to the thickness of the laminations, then an inductance L_o, which is due to the flux induced in these spaces, must be added to L_m to get an expression for the total inductance of a conductor or coil at very high frequencies. The value of L_o will be a fractional part of the inductance of the conductor or coil in air, and its value may be calculated when these distances are known.

The natural frequency of a circuit is

$$f=\frac{1}{2\pi}\sqrt{\frac{1}{(L_m+L_o)C}-\left(\frac{r}{2(L_m+L_o)}-\frac{g}{2C}\right)^2} \tag{14}$$

where r is the resistance of the oscillatory circuit and g is the conductance of the condenser dielectric. The quantities L_o and g will be considered negligible so that (14) becomes

$$f=\frac{1}{2\pi}\sqrt{\frac{1}{L_m C}-\frac{r^2}{4L_m^2}} \tag{15}$$

Solve (15) for L_m,

$$L_m=\frac{1}{4\pi^2 f^2 C}\left(\frac{1+\sqrt{1-(2\pi f r C)^2}}{2}\right) \tag{16}$$

From (12) and (13)

$$L_m=\frac{L_1}{a\,l_o f^{\frac{1}{2}}}\left(\frac{\cosh 2cl_o-\cos 2cl_o}{\cosh 2cl_o+\cos 2cl_o}\right)^{\frac{1}{2}} \tag{17}$$

where $a=\sqrt{0.8\pi^2\lambda\mu 10^{-8}}=0.000281\sqrt{\mu\lambda}$ (18)

For brevity, let $q=2\pi f r C$ (19)

$$\text{"}\quad s=\left(\frac{1+\sqrt{1-q^2}}{2}\right)^{\frac{2}{3}} \tag{20}$$

$$\text{"}\quad v=\left(\frac{\cosh 2cl_o+\cos 2cl_o}{\cosh 2cl_o-\cos 2cl_o}\right)^{\frac{1}{3}} \tag{21}$$

From (16) and (17)

$$\frac{S^{\frac{3}{2}}}{4\pi^2 f^2 C}=\frac{L_1}{a\,l_o f^{\frac{1}{2}} v^{\frac{3}{2}}} \tag{22}$$

from which

$$f=\left(\frac{a\,l_o}{4\pi^2 L_1 C}\right)^{\frac{2}{3}} v\,s \tag{23}$$

The factors v and s are both functions of the frequency but do not vary appreciably from unity, except when the frequency is low or the resistance is high.

For comparatively high frequencies and low resistances, equation (23) becomes

$$f_o = \left(\frac{a\,l_o}{4\,\pi^2 L_1 C}\right)^{\frac{2}{3}} \tag{24}$$

The values of v given in Table I were calculated from assigned values of $2\,cl_o$ in equation (21)

From (9) and (18),

$$2\,c\,l_o = a\,l_o\sqrt{2f} \tag{25}$$

When $2\,cl_o$ is greater than 1.4, the value of f in (25) should be determined by equation (24) but when $2\,cl_o$ is less than 1.4, the value of f to be substituted in (25) is more accurately determined by

$$f_1 = \frac{1}{2\,\pi}\sqrt{\frac{1}{L_1 C} - \frac{r^2}{4\,L_1^2}} \tag{26}$$

where r can generally be considered negligible.

From (23) and (26), (assuming $r=0$ and therefore $s=1$),

$$\frac{f_1}{f} = \sqrt{\frac{\sqrt{2}}{2\,c\,l_o\,v^{\frac{3}{2}}}} \tag{27}$$

Substituting in (27) the values of 1.4 and 1.054 of Table I for $2\,cl_o$ and v respectively shows that f_1 cannot be less than $0.966\,f$ when $2\,cl_o$ is equal to or less than 1.4. Table I shows that f_o (equation (24)) cannot differ from f (equation (23)) more than 8.5 per cent when $2\,cl_o$ is greater than 1.4 and this occurs when $2\,cl_o = 2.4$. It is interesting to note that v varies most from unity when $\tanh 2\,c\,l_o + \tan 2\,cl_o = 0$ which occurs in the table when $2\,cl_o = 0$, 2.4 and 5.5.

The resistance of an oscillatory circuit does not appreciably affect its natural frequency unless the resistance is large. The value of s approaches unity as the value of q approaches zero. Equation (19) shows that $q=0$ when $r=0$ and also that $q=0$ when r is so large that the circuit becomes non-oscillatory since then $f=0$. There must therefore be a value of r for which q is a maximum.

From (15) and (18)

$$q = r\sqrt{\frac{C}{L_m} - \left(\frac{r}{2}\cdot\frac{C}{L_m}\right)^2} \tag{28}$$

which gives a maximum value of q when

$$r = \sqrt{\frac{2\,L_m}{C}} \tag{29}$$

TABLE I

$2cl_o=$	0.2	0.4	0.6	.8	1.0	1.2	1.4	1.6	1.8	2.0
$v=$	3.6832	2.322	1.774	1.470	1.277	1.145	1.054	0.9925	0.9524	0.9286
$2cl_o=$	2.2	2.4	2.6	2.8	3.0	3.2	3.4	3.6	3.8	4.0
$v=$	0.9173	0.9149	0.9187	0.9264	0.9363	0.9472	0.9579	0.9678	0.9767	0.9842
$2cl_o=$	4.2	4.4	4.6	4.8	5.0	5.2	5.4	5.6	5.8	6.0
$v=$	0.9903	0.9950	0.9985	1.0010	1.0026	1.0035	1.0038	1.0038	1.0036	1.0032

TABLE II

$q=$	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
$s=$	0.9983	0.9933	0.9845	0.9719	0.9548	0.9322	0.9022	0.8618	0.8018	0.6300

which may be compared with the resistance $r=\sqrt{\frac{4L_m}{C}}$, which renders the circuit non-oscillatory.

Substituting (29) in (28) shows that

$$q_{max.}=1 \tag{30}$$

The value of q depends upon the product $r\,f$, and f decreases as r increases, therefore q and s do not vary greatly with changes in the value of r. Table II gives values for q and the corresponding values of s as computed by means of equation (20).

The determination of q is made in a manner similar to $2\ cl_o$ The value of f to be substituted in (19) is computed from either (24) or (26). When both v and s have been determined, the natural frequency of the circuit can be evaluated by means of equation (23). If $v\,s$ differs much from unity, then the value of f in (23) can be used to determine new values of $2\ cl_0$ and q from which v and s will give a more accurate solution for f in (23).

The value of L_1 in (23) and (24) can be calculated from the equation

$$L_1=\frac{4\,\pi\,\mu\,A\,N^2}{d} \tag{31}$$

for a coil having a magnetic circuit the mean length of which $=d$, cross-sectional area of iron core $=A$, and a total of N turns of wire forming a layer over a large part of the length d.

To illustrate the practical application of equation (23) an example will be given.

Assume $d=50$ cm., $l_o=0.03$ cm., $\lambda=1.2\times10^4$, $\mu=2000$, $A=100$ sq. cm., $N=20$, and $C=0.1$ microfarad.

Then from (28)

$$L_1=\frac{4\,\pi\times2000\times100\times400}{50}=6.4\,\pi\times10^6 \text{ cm.}$$

Substitute in (18)

$$a\,l_o=0.000281\times0.03\times\sqrt{2000\times1.2\times10^4}=0.0413$$

From these values in (24)

$$f_o=\left(\frac{0.0413}{4\,\pi^2\times6.4\,\pi\times10^6\times0.1\times10^{-15}}\right)^{\frac{2}{3}}=8{,}793 \text{ cycles.}$$

To find the value of v substitute in (25), which gives

$$2\,c\,l_o=0.0413\sqrt{2\times8793}=5.477$$

Referring to Table I, $v=1.0038$ when $2\ c\,l_o=5.477$. This correction factor adds 33 cycles to f_o or

$$f=f_o\,v=8{,}826 \text{ cycles.}$$

The natural frequency of the circuit, if L_1 is substituted for L_m (equation (26)), is

$$f_1 = 3{,}550 \text{ cycles},$$

or the presence of eddy currents in the iron increases the natural frequency by the factor 2.5.

The circuit was considered to have negligible resistance. A resistance of 45 ohms is required to decrease the frequency one per cent. It is interesting to determine the resistance which will give the maximum value of q.

Substitute $f_o\, s$ for f in (25) which is equivalent to multiplying the former value of

$2\,c\,l_o$ by $\sqrt{s}$, or $2\,c\,l_o = 5.477\,\sqrt{0.63} = 4.35$ or $v = 0.9938$, therefore

$$f = 8793 \times 0.63 \times 0.9938 = 5{,}505 \text{ cycles}.$$

In (19), put $q = 1$ and $f = 5{,}505$, and solve for r; then $r = 289$ ohms.

It will be noted that the quantity L_m may be defined by equations (16) and (17), or by the equation,

$$\tfrac{1}{2} L_m I^2 = \tfrac{1}{2} C E^2 \qquad (32)$$

in which, referring to Figure 1, I is the initial current, E is the potential to which the capacitance C becomes charged, the resistance of the circuit is considered negligible, and the potential of the battery is considered negligible compared to the potential E. If the value of L_m is considered the same in both instances, then the potential E can be calculated. A consequence of this assumption is that the eddy current loss, W, during the first quarter oscillation is

$$W = \tfrac{1}{2} I^2 (L_1 - L_m) \qquad (33)$$

where $\frac{1}{2} L_1 I^2$ is the electromagnetic energy stored in the coil when the switch S is opened. The voltage E may be assumed to rise sinusoidally, from which the speed required in opening the switch S without a spark can be estimated.

The effective inductance in the direct current circuit of a Poulsen arc generator depends upon the natural frequency of the magnet coils and generator in series with the antenna capacitance, which form an oscillatory circuit during the interval of each cycle of the radio frequency current that the arc is extinguished. A large loss occurs in the iron due to radio frequency eddy currents.

Increase of permeance due to the presence of iron is accompanied by an increase of conductance. This inherent property of iron will not allow its use in the magnetic circuit of a radio frequency oscillatory electric circuit where efficiency is an essential requirement.

SUMMARY: In radio or high audio frequency circuits containing iron-core inductances, the eddy currents induced in the iron cause a marked change of effective inductance which is dependent on the frequency.

Following the Steinmetz procedure, the author finds the magnetic flux density in a laminated iron core with alternating current excitation. He then derives expressions and tables for determining the natural frequency of circuits containing iron-core inductances. The results obtained are numerically illustrated.

DISCUSSIONS

on

'THE ELECTRICAL OPERATION AND MECHANICAL DESIGN OF AN IMPULSE EXCITATION MULTI-SPARK GROUP RADIO TRANSMITTER'† BY BOWDEN WASHINGTON

FIRST DISCUSSION*

By

LIEUTENANT ELLERY W. STONE, U. S. N. R. F.

(DISTRICT COMMUNICATION SUPERINTENDENT, NAVAL COMMUNICATION SERVICE, SAN DIEGO, CALIFORNIA)

Mr. Washington's paper is an excellent contribution to a subject which has been of considerable interest to me for the last four years. However, I cannot subscribe to his statement in the first paragraph of his article that impact excitation has not "been put into thoroly practical operation before." To do so is to ignore the published accounts of various other types of prior impulse excitation transmitters.

In particular, I have reference to the impulse excitation transmitter of the C. Lorenz Company of Berlin, described in volume 1, number 4, of the PROCEEDINGS of this Institute, the impulse excitation transmitter described by me in volume 4, number 3 and volume 5, number 2, of the PROCEEDINGS, and the impulse transmitters of the Kilbourne & Clark Manufacturing Company which have been described in United States letters patent owned by that company and in other technical publications.

The extensive use and thoroly practical features of the "Multitone" (Lorenz) system are too well known to require further discussion here.

In connection with the impulse excitation transmitter designed by me, I may state that following the publication of my gap and gap circuit designs in the two papers noted above, with a view toward allowing any interested party to use them gratis, this system of transmitter was adopted by the Haller Cunningham

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† Published in the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, number 6.

Electric Company of San Francisco. Certain mechanical improvements in the original design have been made by them from time to time. The practical adoption which has been accorded this type of transmitter is evidenced by that company's advertisements, appearing monthly in the PROCEEDINGS.

The Kilbourne & Clark Company, who were pioneers in this work, have on the market two types of impulse transmitters, known respectively as the "Simpson" and "Thompson" transmitters (after their designers). A novel type of gap together with a gap circuit of aperiodic constants is employed in each. It does not appear necessary, in the face of their extensive sales, to comment on the practicability of their operation.

It may be of interest to note that the first impulse excitation transmitter was designed by Lodge in his United States Letter Patent 609,154, inasmuch as he provided for gap and antenna circuits without resonant tuning, and for a gap of high resistance ("polished and protected from ultra-violet light, so as to supply the electric charge in as sudden a manner as possible"). With the expiration of this patent in 1915, the field was opened for its adoption by those companies not desiring to make use of the Marconi four-tuned-circuit patent.

In this connection, interest attaches itself to a statement of somewhat prophetic tone made by Mr. Robert Marriott in 1913, in his discussion of the Lorenz paper noted above, which I quote below:

"It is interesting to note that in the case of 'ideal' impulse excitation, where the primary and secondary circuits need not be syntonized, various patents covering such tuning are avoided."

The use of an untuned secondary receiver with an impulse transmitter, which procedure has been adopted by Kilbourne & Clark, Haller Cunningham, and apparently by Cutting and Washington, is a further reversion to the disclosures of the Lodge patent, in which such a receiver is described. I am in hearty agreement with Mr. Washington as to the advantages of this type of receiver over the coupled tuned circuit type, when used commercially with a crystal detector.

The almost undamped or continuous nature of the antenna current due to partial discharges, noted by Mr. Washington, is characteristic of true impulse excitation, as observed in my papers in which comment was made on the increase in signal strength with tikker and beat reception.

Mr. Washington's "concentration circuit" appears to play the same rôle as the "tone" circuit of other impulse transmitters.

SECOND DISCUSSION

(In Answer to the Preceding Discussion)

By

Ensign Bowden Washington, U. S. N. R. F.

I would like to say in answer to Lieutenant Stone's interesting criticism of my statement, that to the best of my knowledge, "Impact excitation has not been put in a thoroly practical operation before." I was speaking only from such data as I had available on this subject, and it is more than possible that my remarks were too broad in scope. The inapplicability of Lieutenant Stone's criticism appears to be somewhat dependent on the definition of "impact excitation." I have always felt that this expression should be taken to mean *no oscillations in the primary*, that is, the transfer of energy should occur during one current pulse. Otherwise it is only a matter of degree, and this seems objectionable for a definition.

It was my impression, gathered in Europe in 1913, that very few Lorenz, or "Multitone," transmitters were put in practical commercial operation, and that these few gave considerable trouble. Among other things mentioned were commutator troubles, usually present in high voltage direct current machines; and it was stated that the tone circuit was extremely critical to the condition of the gap. I found this latter to be true of the original Chaffee gap. I also have been told by British and French signal officers that the German Army purchased a considerable number of these sets which had to be discarded later as highly unreliable.

I am strongly of the impression that the standard Kilbourne & Clark transmitter, as marketed, is not true impact excitation, but is highly quenched. I had an opportunity to make quite extensive tests on a 2 kilowatt commercial set of their manufacture at the Cruft Laboratory of Harvard University, using a Braun tube in these tests. I was never able to get less than 2.5 complete oscillations in the primary circuit, and the set displayed one characteristic which strongly prejudiced me towards the belief that it was not operating as a true impact set, apart from the evidence obtained with the Braun tube. It is extremely critical as to tuning, more so, it seems, than the average quenched set. In this connection I would like to state that I have seen some Braun tubes oscillograms taken on this type of transmitter

purporting to show impact excitation. These were taken, I believe, of the primary condenser voltage with no time axis deflection. A single heavy line on one side of the zero was the result, supposedly showing potential in but one direction. This can be easily explained;—as the condenser charges the spot moves very slowly, allowing the fluorescent screen to be exposed to the spot for some time. The actual oscillations occur in a very much shorter time, and unless every effort is made to have the spot exceedingly bright, there will be no lighting of the path of the spot due to the oscillations. The gaps appeared very little different in general construction from the ordinary quenched gap, except for what seemed rather inadequate cooling surfaces, and the gap length, 0.010 inch (0.4 mm.), seems long for this type of excitation.

The writer did considerable work on the "Hytone" gap, as manufactured by the Clapp Eastham Company, at the Jefferson Physical Laboratory, at Harvard University, with Dr. Chaffee in the spring of 1914. The Braun tube was used continuously; and on direct current, using a smooth disk it was possible to obtain pure impact excitation under ideal conditions. With a segmented gap, the following difficulty was encountered both on direct and alternating current. When the gap was opened, that is, when the faces of the segments were not opposite to each other, the condenser was given an opportunity to charge to considerable potential. On the approach of the segments a long, stringy spark would pass, followed by several small discharges. The only way to obtain the energy of this first spark appeared to be to tune for resonance, and adjust for proper coupling. It was, however, possible to quench entirely the succeeding minor sparks by properly proportioning the circuits, but in either case some energy was lost, and the apparatus was far from efficient. Alcohol vapor was tried, but never pressures of any real magnitude. Sixty cycle alternating current and a tone circuit were also tried on the smooth gap, but both the tone quality and tone efficiency seemed poor. The only thoroly satisfactory combination as to efficiency and functioning appeared to be the smooth gap, direct current, and a tone circuit. This was open to the practical objections of a high voltage direct current generator. Also, tho several gaps were constructed, it would seem difficult to obtain a gap which would keep its adjustments for spark length thruout the range of temperature encountered during a long run starting with a cold gap.

I should like to take exception to the term "partial dis-

charges." I cannot conceive of a discharging condenser being stopped in "mid air," so to speak, and if it is meant that the condenser is only partially charged, what constitutes a full charge?

It may also be pointed out that the "concentration circuit" does not function in the same manner, or rather to the same purpose, as a tone circuit. This type of transmitter is entirely dependent on the generator frequency as to the radio frequency of the emitted groups. The concentration circuit is merely to group the discharges so as to give the proper form to the antenna envelope for maximum tone efficiency.

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PROCEEDINGS

of

The Institute of Radio Engineers

(INCORPORATED)

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EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
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ANNUAL AWARD OF A MEDAL OF HONOR

BY THE BOARD OF DIRECTION OF THE INSTITUTE

By vote of the Board of Direction at the February 15, 1917, meeting, there will be awarded annually a Medal of Honor of The Institute of Radio Engineers under the conditions set forth below. The medal, which has been designed by the well-known sculptor, Mr. Edward Field Sanford, Junior, of New York, has on its face a symbolic representation of electromagnetic waves and the words: "INSTITUTE OF RADIO ENGINEERS." On the reverse are a laurel wreath and the words: "To ——— IN RECOGNITION OF DISTINGUISHED SERVICE IN RADIO COMMUNICATION ——— (date)."

The medal will be awarded annually by vote of the Board of Direction of the Institute at its April meeting to that person, who, during the previous year (January thru December) shall have made public the greatest advance in the science or art of radio communication. The advance may be an unpatented or patented invention, but it must be completely and adequately described in a scientific or engineering publication of recognized standing and must be in actual (tho not necessarily commercial) operation. However, preference will be given to widely used and widely useful inventions. The advance may also be a scientific analysis or explanation or hitherto unexplained phenomena of distinct importance to the radio art, tho the application thereof need not be immediate. Preference will be given analyses directly applicable in the art. In this case, also, publication must be in full and approved form. The advance may further be a new system of traffic regulation or control; a new system of administration of radio companies or of the radio service of steamship, railroad, or other companies; a legislative program beneficial to the radio art; or any portion of the operating or regulating features of the art. It must be publicly described in clear and approved form, and must, in general, be actually adopted in practice.

The method of awarding the medal is as follows:

1. At least 30 days before the April meeting, the Board of Direction shall call, from a number of Members and Fellows

of the Institute whom it may choose to consult, for suggested candidates.

2. At the April meeting of the Board, those actually present or voting by mail shall nominate at least one, but not more than three candidates for the award, in order of preference. The names of these candidates shall then be sent to each member of the Board, and each member of the Board shall have the privilege of returning a vote for one candidate. Four weeks after the April meeting, the ballots shall be read, and the candidate receiving the most votes shall be awarded the medal.

3. The official presentation of the medal to the successful candidate or his representative shall be at the May or June meeting immediately following. The person awarded the medal shall be privileged to indicate this fact in giving his titles and honors in the fashion customary in learned and artistic societies: thus—Mr. William Jones, E.E., Medal (or Award) of Honor, Institute of Radio Engineers, 191–.

For 1917, the medal was awarded to Mr. (now Captain) Edwin H. Armstrong in recognition of his work and publications dealing with the action of the oscillating and non-oscillating audion. For 1918, in view of the limitation of publication brought about by war conditions, no award was made.

THE MORRIS LIEBMANN MEMORIAL PRIZE

The Board of Direction of The Institute of Radio Engineers, at its regular meeting held on February 5, 1919, accepted a gift of $10,000 to The Institute of Radio Engineers, from an anonymous donor, himself a friend and member of the Institute, to "preserve the memory of our late friend and fellow member, Colonel Morris N. Liebmann, who has sacrificed his life in the cause of our country."

The principal of this fund will be preserved in perpetuity and the annual income derived therefrom only will be expended. The present amount of this income is $425.00 per annum, and is to be awarded each year on the first day of October (beginning October 1, 1919), by a special committee appointed annually by the Board of Direction, to that member of the Institute, who, in the opinion of this committee, shall have made the most important contribution to the radio art during the preceding calendar year.

This annual award is to be known as the "Morris Liebmann Memorial Prize" and it is hoped will act as an additional incentive to the further rapid development of radio communication.

THEORY AND OPERATING CHARACTERISTICS OF THE THERMIONIC AMPLIFIER*

By

H. J. van der Bijl, M.A., Ph.D.

(Western Electric Company, Incorporated, New York City)

I. Introduction

The three-electrode thermionic tube has been responsible for a great deal of the recent rather remarkable developments in the art of radio communication. In its most commonly known form it consists of an evacuated vessel containing a hot filament cathode, an anode placed at a convenient distance from the cathode and a third electrode in the form of a grid placed between cathode and anode. To discuss in detail the theory of operation of the device in its various applications, such as oscillation generator, radio detector, and amplifier would be beyond the scope of the present paper. What I intend to give here is merely its fundamental principles of operation, with particular reference to its application as an amplifier. The framework of this theory was worked out in the winter of 1913-14 and formed the basis of a considerable amount of research and development work that has since been done in this laboratory on the device and its various applications.

A condition which is assumed in the elaboration of the views expressed in the following is that the operation of the device is independent of any gas ionization, or in other words, that the current is carried almost entirely by the electrons emitted from the hot cathode. It is, of course, to be understood that it is at present impossible completely to eliminate ionization by collision of the electrons emitted from the cathode with the residual gas molecules. But the condition assumed can always be realized practically by evacuating the tube to such an extent that the number of positive ions formed by collision ionization is always small compared with the number of electrons moving from cathode to anode. This happens when the mean free path of the electrons in the residual gas becomes large compared with

* Received by the Editor, October 28, 1918.

the dimensions of the device. The pressure necessary for this is not very low, and were it not for the gases occluded in the electrodes and walls of the vessel, it would be a comparatively simple matter to make the tube operate independently of gas ionization. The energy liberated by the electrons striking the anode, however, usually causes a sufficient rise in the temperature of the device to liberate enough gas to increase the pressure unduly.[1]

This is especially marked in the case of tubes handling large amounts of power. It is, therefore, necessary to denude the electrodes and walls of the tube of gases during the process of evacuation. Furthermore, since the energy liberated at the anode increases with the applied voltage, it is seen that this voltage must be kept within limits depending upon the degree of evacuation obtained. This is very important when using the device as a telephone relay, as was recognized by Dr. Arnold of this laboratory in the early stages of his experiments with this type of device. As is well known to workers in this field, it is difficult to keep a discharge steady and reproducible when ionization by collision is appreciable, and steadiness and reproductibility are conditions which must be complied with by a telephone relay.

The success of the tubes developed by the Western Electric Company is mainly due to the extensive study that has been made of the bearing of the structural parameters of the device on its operation. It is hardly possible to meet the requirements of efficiency and satisfactory operation of the device without an explicit mathematical formulation of its operation. A satisfactory telephone relay must, for example, do more than merely utilize the direct current power in its local circuit to amplify alternating current power: it must faithfully reproduce the incoming speech currents, it must also be capable of handling sufficient power, and have a definite impedance that can conveniently be made to fit the impedance of the telephone line. Since all these conditions depend on the structural parameters of the amplifier, they will not be satisfied unless the amplifier be properly designed, and so much distortion may be produced as to make the device worthless as a telephone relay. On the other hand, it has been found that by properly designing the amplifier the above-named requirements can be met very satisfactorily.

[1] For a fuller explanation of the effect of gas, see H. J. van der Bijl, "Phys. Rev.", (2), 12, page 174, 1918.

II. Current-Voltage Characteristics of Simple Thermionic Devices

We shall not here enter into a discussion of the extensive investigations that have been carried out on thermionics, but merely, for the purpose of elucidation, touch upon those phases of the subject which have a direct bearing on the theory of operation of the thermionic amplifier.

Consider a structure consisting of a heated cathode and an anode, and contained in a vessel which is evacuated to such an extent that the residual gas does not play any part in the current convection from cathode to anode. The number of electrons emitted from the cathode is a function of its temperature. If all the electrons emitted from the cathode pass to the anode, the relation between the resulting current I and cathode temperature T is given by a curve of the nature shown in Figure 1. This curve is obtained provided the voltage between anode and cathode is always high enough to drag all the electrons to the anode as fast as they are emitted from the cathode; that is, I in Figure 1 represents the saturation current. The saturation

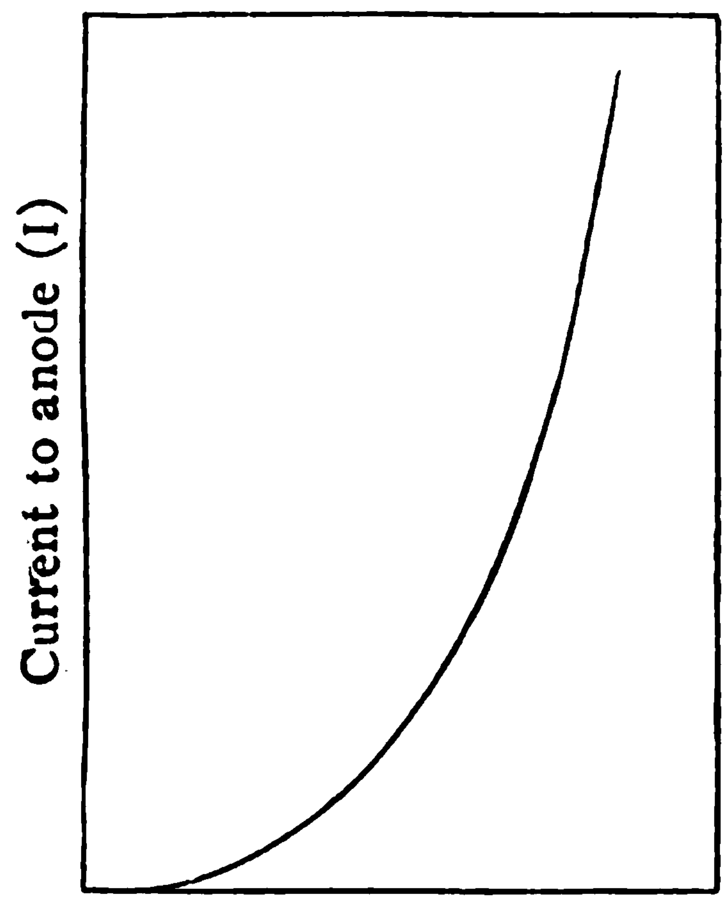

FIGURE 1

current is obtained in the following way: Suppose the cathode be maintained at a constant temperature T_1 and the voltage V between anode and cathode be varied. As this voltage V is raised from zero, the current I to the anode at first increases,

the relation between V and I being represented by the curve OA_1 of Figure 2. Any increase in V beyond the value corresponding to A_1 causes no further increase in I, and we get the part A_1B_1 of the curve. Clearly this part of the curve corresponds

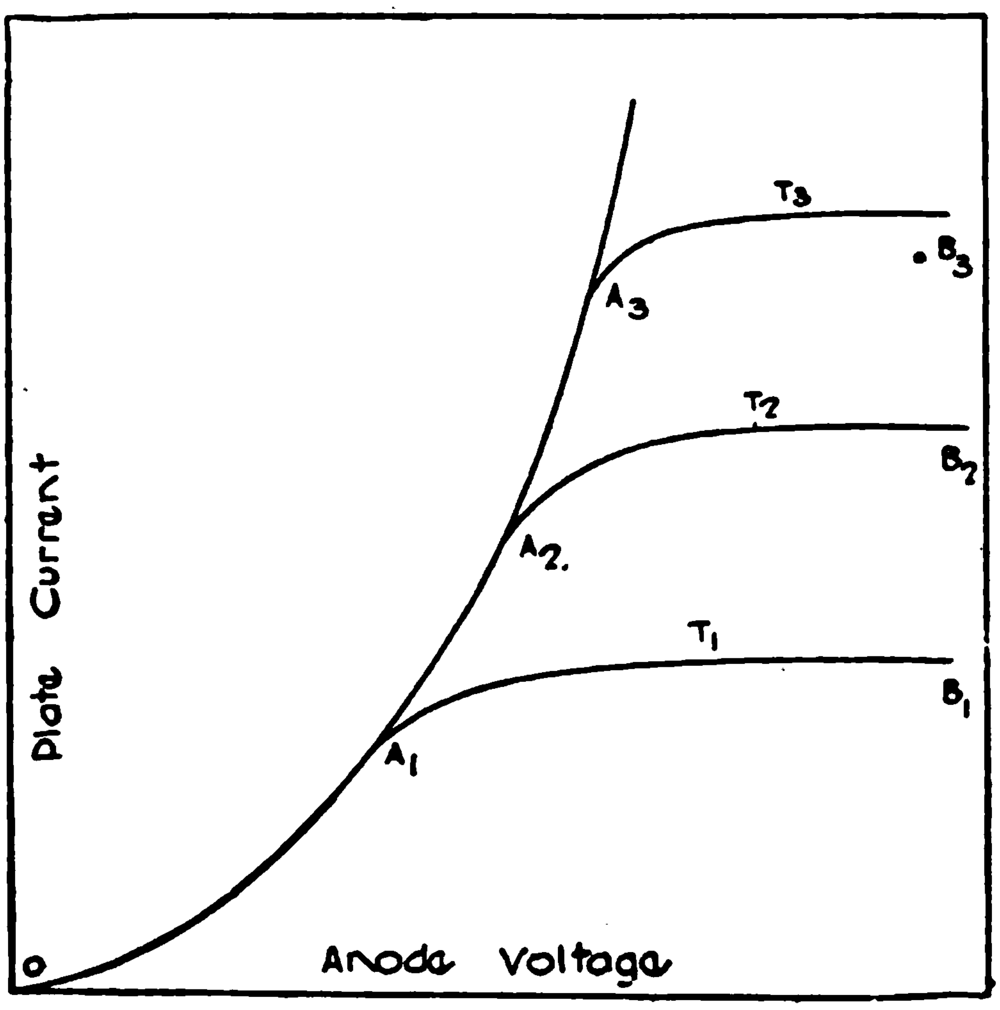

FIGURE 2

to the condition when all the emitted electrons are drawn to the anode as fast as they are emitted from the cathode. If the cathode temperature be increased to T_2, the number of emitted electrons is increased, and we get the curve OA_2B_2. When these values of the saturation current are plotted as a function of the temperature, we obtain the curve of Figure 1. This curve is represented very approximately by the equation

$$I = aT^{\frac{1}{2}}e^{\frac{b}{T}}, \tag{1}$$

where a and b are constants. This equation was derived by O. W. Richardson in 1901[2] on the basis of the theory that the electrons are emitted from the hot cathode without the help of any gas, but solely in virtue of their kinetic energy. The formulation of this theory was the first definite expression of what may be termed a pure electron emission.

In the state in which Richardson's equation holds the cur-

[2] "Proc. Camb. Phil. Soc.," volume II, 285, 1901; "Phil. Trans. Roy. Soc.," A, 201, 1903.

rent is independent of the voltage. Under these conditions the device to be treated in the following does not function as amplifier or detector, since it depends for its operation on the variation of current produced by variation of the voltage. In this device the current established in the circuit connecting filament and anode (that is, the so-called output circuit) by the electrons flowing from filament to anode is varied by potential variations applied between the filament and grid. The condition under which the current is a function of the voltage is represented by the part OA of Figure 2. Here the voltage is not high enough to draw all the electrons to the anode as fast as they are emitted from the cathode; in other words, there are more electrons in the neighborhood of the cathode than can be drawn away by the applied voltage. It was first pointed out explicitly by C. D. Child in 1911 that this limitation to the current is due to the space charge effect of the electrons in the space between anode and cathode. The influence of space charge is something which must always be considered where conduction takes place by means of dislodged electrons or ions, such as the conduction thru gases at all pressures, liquids, and high vacua. Assuming that in the space only ions of one sign are present, Child deduced the equation:[3]

$$I = \frac{1}{9\pi}\sqrt{\frac{2e}{m}} \cdot \frac{V^{\frac{3}{2}}}{x^2} \tag{2}$$

In this equation, which was deduced on the assumption that both cathode and anode are equipotential surfaces of infinite extent, I is the thermionic current per square centimeter of cathode surface, V the voltage between anode and cathode, x the distance between them, and e and m the charge and mass of the ion, respectively.

When the full space charge effect exists, the current is independent of the temperature of the cathode. This can be understood more easily with reference to Figure 3, which gives the current as a function of the temperature of the cathode for various values of the voltage between anode and cathode. Suppose a constant voltage V_1 be applied between anode and cathode, and the temperature of the cathode be gradually increased. At first when the temperature is still low, the voltage V_1 is large

[3] C. D. Child, "Phys. Rev.," 32, 498, 1911. The space effect has been fully studied by J. Lilienfeld ("Ann. d. Phys.," 32, 673, 1910); I. Langmuir ("Phys. Rev.", (2), 2, 450, 1913), who also independently derived the space charge equation (2) and published a clear explanation of the limitation of current by the space charge; and Schottky ("Jahrb. d. Rad. u. Elektronik," volume 12, 147, 1915).

enough to draw all the emitted electrons to the anode, and an increase in the temperature results in an increase in the current. This gives the part OC_1 of the curve of Figure 3. When the temperature corresponding to C_1 is reached, so many electrons are emitted that the resulting volume density of their charge causes all other emitted electrons to be repelled, and these return to the filament. Obviously any further increase in the tempera-

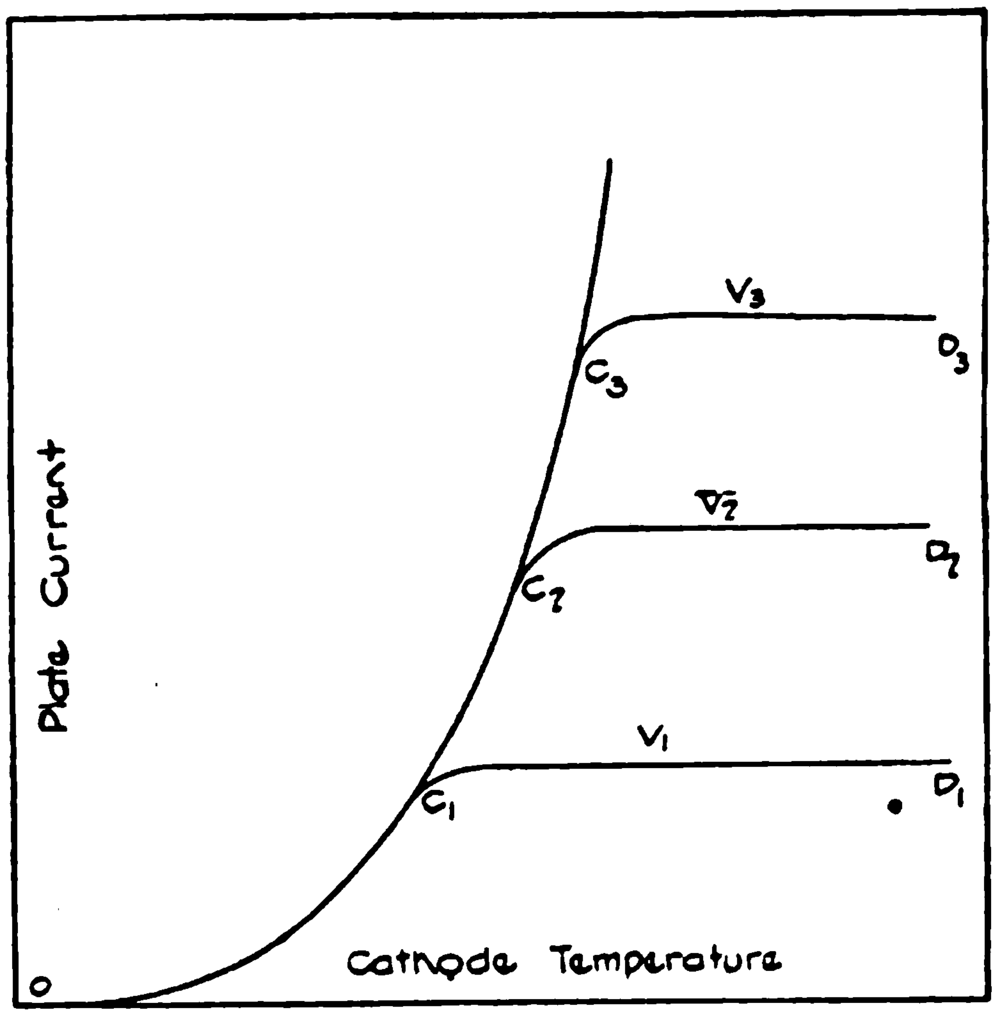

FIGURE 3

ture of the cathode beyond that given by C_1 causes no further increase in the current, and we obtain the horizontal part C_1D_1. If, however, the voltage be raised to V_2, the current increases, since more electrons are now drawn away from the supply at the filament, the full space charge effect being maintained by less emitted electrons being compelled to return to the filament. It is now clear that the part OC of Figure 3 corresponds to the part AB of Figure 2 and CD of Figure 3 to OA of Figure 2. The latter represents the condition under which the thermionic amplifier operates.

It is important to note that the thermionic amplifier operates under the condition characterized by the circumstances that the applied voltage is not sufficiently high to give the saturation current.

III. Action of the Auxiliary Electrode

So far we have considered the case of a simple thermionic device consisting of a cathode and anode. When a third electrode is added to the system, the matter becomes more complicated.

The insertion of a third electrode to control the current between cathode and anode is due to de Forest.[4] De Forest later gave this electrode the form of a grid placed between cathode and anode.[5] About the same time von Baeyer[6] used an auxiliary electrode in the form of a wire gauze to control thermionic discharge. The gauze was placed between the thermionic cathode and the anode.

The quantitative effect of the auxiliary electrode was first given by the present writer.[7]

To get an idea of the effect of the auxiliary electrode consider the circuit shown in Figure 4. *F* denotes the cathode, *P* the anode, and *G* the auxiliary electrode which is in the form of a grid between *F* and *P*. Let the potential of *F* be zero, and that of *P* be maintained positive by the battery *E*, and let E_c for the present be zero. Now, altho there is no potential difference between *F* and *G*, the electric field between *F* and *G* is not zero,

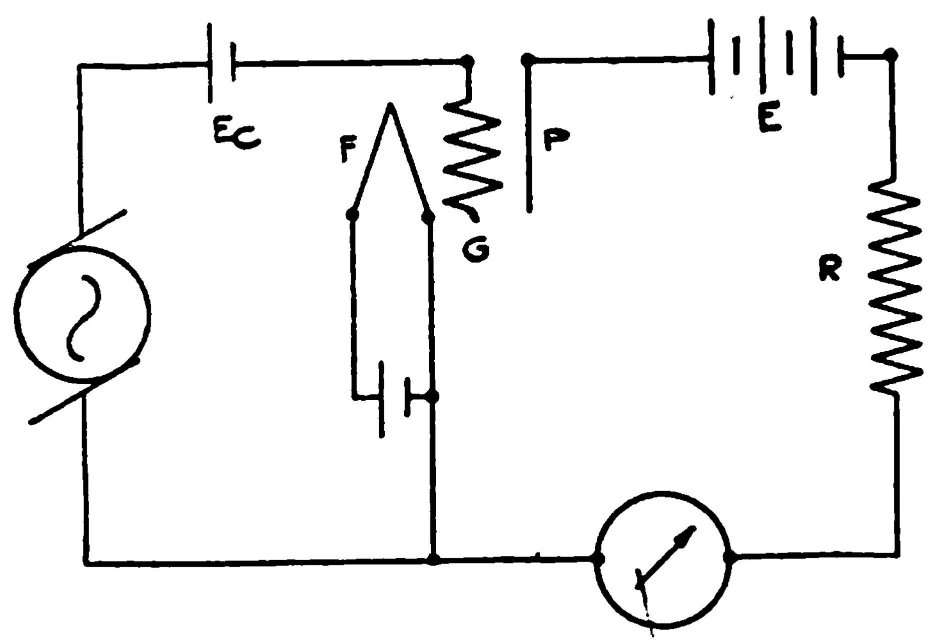

Figure 4

[4] De Forest, U. S. Patent number 841,387, 1907.
[5] De Forest, U. S. Patent number 879,532, 1908.
[6] von Baeyer, "Verh. d. D. Phys. Ges.," 7, 109, 1908.
[7] H. J. van der Bijl, "Verh. d. D. Phys. Ges.," May, 1913, page 338. In these experiments which were also performed under such conditions that the current was carried almost entirely by electrons, the source of electrons was a zinc plate subjected to the action of ultra-violet rays. It is obvious that the action of the auxiliary electrode is independent of the nature of the electron source. Hence the results then found apply also to the present case.

but has a finite value which depends upon the potential of P. This is due to the fact that the potential of P causes a stray field to act thru the openings of the grid. If the potential of P be E_B the field at a point near F is equal to the field which would be sustained at that point if a potential difference equal to γE_B were applied directly between F and an imaginary plane coincident with the plane of G, where γ is a constant which depends on the mesh and position of the grid. If the grid is of very fine mesh γ is nearly zero, and if the grid be removed—that is, if we have the case of a simple valve—γ is equal to unity.

These results can be expressed by the following equation:

$$E_s = \gamma E_B + \varepsilon. \tag{3}$$

Here ε is a small quantity which depends upon a number of factors, such as the contact potential difference between cathode and grid and the power developed in the filament, which is the usual form of cathode used. It is generally of the order of a volt and can be neglected when it is small compared with γE_B. Obviously the current between anode and cathode depends on the value of E_s.

Now, suppose a potential E_c be applied directly to the grid G, the cathode F remaining at zero potential. The current is now a function of both E_s and E_c':

$$I = \Phi\,(E_s, E_c). \tag{4}$$

Before determining the form of this function, let us consider in a general way how the current is affected by E_s and E_c. We have seen that E_s is due to the voltage E_B between anode and cathode, and is less than E_B if the grid is between anode and cathode, since in this case γ is always less than unity. Under the influence of E_s the electrons are drawn thru the openings of the grid and are thrown on to the anode by the strong field existing between grid and anode. The effect of E_c on the motion of the electrons between F and G is similar to that of E_s. Whether or not electrons will be drawn away from the cathode depends on the resultant value of E_s and E_c. If E_s+E_c is positive, electrons will flow away from the cathode, and if E_s+E_c is zero or negative, all the emitted electrons will be returned to the cathode, and the current thru the tube will be reduced to zero. Now, E_s+E_c will be positive: (1) when E_c is positive (E_s is always positive), and (2) if E_c is negative and less than E_s.

1. When E_c is positive, some of the electrons moving toward the grid are drawn to the grid, while the rest are drawn thru the openings of the grid to the anode under the influence of E_s.

The relative number of electrons going thru and to the grid depends upon the mesh of the grid, diameter of the grid wire, and the relative values of E_s and E_c. When, for example, E_s is large compared with E_c, the number of electrons going to the grid is comparatively small, but for any fixed value of E_s the current to the grid increases rapidly with increase in E_c. Hence, for positive values of E_c, current will be established in the circuit $F\,G\,E_c$, Figure 4.

2. If, however, E_c is negative and less than E_s, as was the case in the above named experiments of the writer, nearly all the electrons drawn away from the filament pass to the plate, practically none going to the grid. In this case the resistance of the circuit $F\,G\,E_c$ is infinite.

If, now, an alternating emf. be impressed upon the grid so that the grid becomes alternately positive and negative with respect to the cathode, the resistance of the circuit $F\,G\,E_c$, which may be referred to as the input circuit, will be infinite for the negative half cycle and finite and variable for the positive half cycle. If, on the other hand, the alternating emf. be superimposed upon the negative value, E_c, the values of these voltages being so chosen that the resultant potential of the grid is always negative with respect to the cathode, the impedance of the input circuit is always infinite.

Broadly speaking, the operation of the thermionic amplifier is as follows: The current to the anode we have seen is a function of E_s and E_c, or keeping the potential E_B of the anode constant, the current for any particular structure of the device is a function only of the potential on the grid. Hence, if the oscillations to be repeated are impressed upon the input circuit, variations in potential difference are set up between cathode and grid, and these cause variations in the current in the circuit FPR, the power developed in the load R being greater than that fed into the input circuit. It is seen then that the device functions broadly as a relay in that variations in one circuit set up amplified variations in another circuit unilaterally coupled with the former.

IV. Current-Voltage Characteristic of the Thermionic Amplifier

Equation (2) which gives the current to the anode as a function of the applied voltage in the case of a simple device containing equipotential electrodes of infinite extent is of little use in deriving the amplification equations of the thermionic ampli-

fier. In the first place, the cathode in this device is not an equipotential surface, but a filament which is heated by passing a current thru it. Secondly, the insertion of a grid between the filament and the anode so complicates the electric field distribution that a theoretical deduction of the relation between the current to the anode and the applied voltages between filament and grid and filament and anode is difficult and leads to expressions that are too complicated for practical use. I have, therefore, found it more practical to determine the characteristic of the tube empirically, and found as the result of a large number of experiments that the characteristic can be represented with sufficient accuracy by the following equation:

$$I = a\,(E_s + E_c)^2, \tag{5}$$

where a is a constant depending on the structure of the device.[8]

With the help of equation (3) this becomes

$$I = a\,(\gamma E_B + E_c + \varepsilon)^2. \tag{6}$$

This gives the current to the anode as a function of the anode and grid potentials, the potential of the filament being zero. If a number of voltages be impressed upon the grid and anode, we have generally

$$I = a\,(\gamma \Sigma E_B + \Sigma E_c + \varepsilon)^2. \tag{7}$$

If, for example, an alternating emf., $e \sin p\,t$ be superimposed upon the grid-voltage, E_c, the equation becomes:

$$I = a\,(\gamma E_B + E_c + e \sin p\,t + \varepsilon)^2. \tag{8}$$

It must be understood that equation (6) gives the direct characteristic of the device itself; that is, E_B in equation (6) is the voltage directly between the filament and the anode P (Figure 4). If the resistance R be zero, E_B is always equal to E, the voltage of the battery in the circuit $EPRE$, which is constant. If R be not zero, the potential difference established between the ends of R by the current flowing in it makes E_B a function of the current. The effect of the resistance R on the characteristic will be explained later. For the present we shall confine ourselves to a discussion of the characteristic of the amplifier itself. This characteristic can always be obtained experimentally by making R equal to zero and using an ammeter in the circuit $FPER$ (Figure 4), the resistance of which is small compared with the internal output resistance of the amplifier itself.

[8] Altho this equation is sufficiently accurate when using the device as an amplifier, its accuracy does not suffice for purposes of detection, since the detection action is a function of the second derivative of the characteristic.

A graphical representation of equation (6) is given in Figure 5. The curves give the current to the anode as a function of the grid voltage E_c for different values of the parameter, E_B.

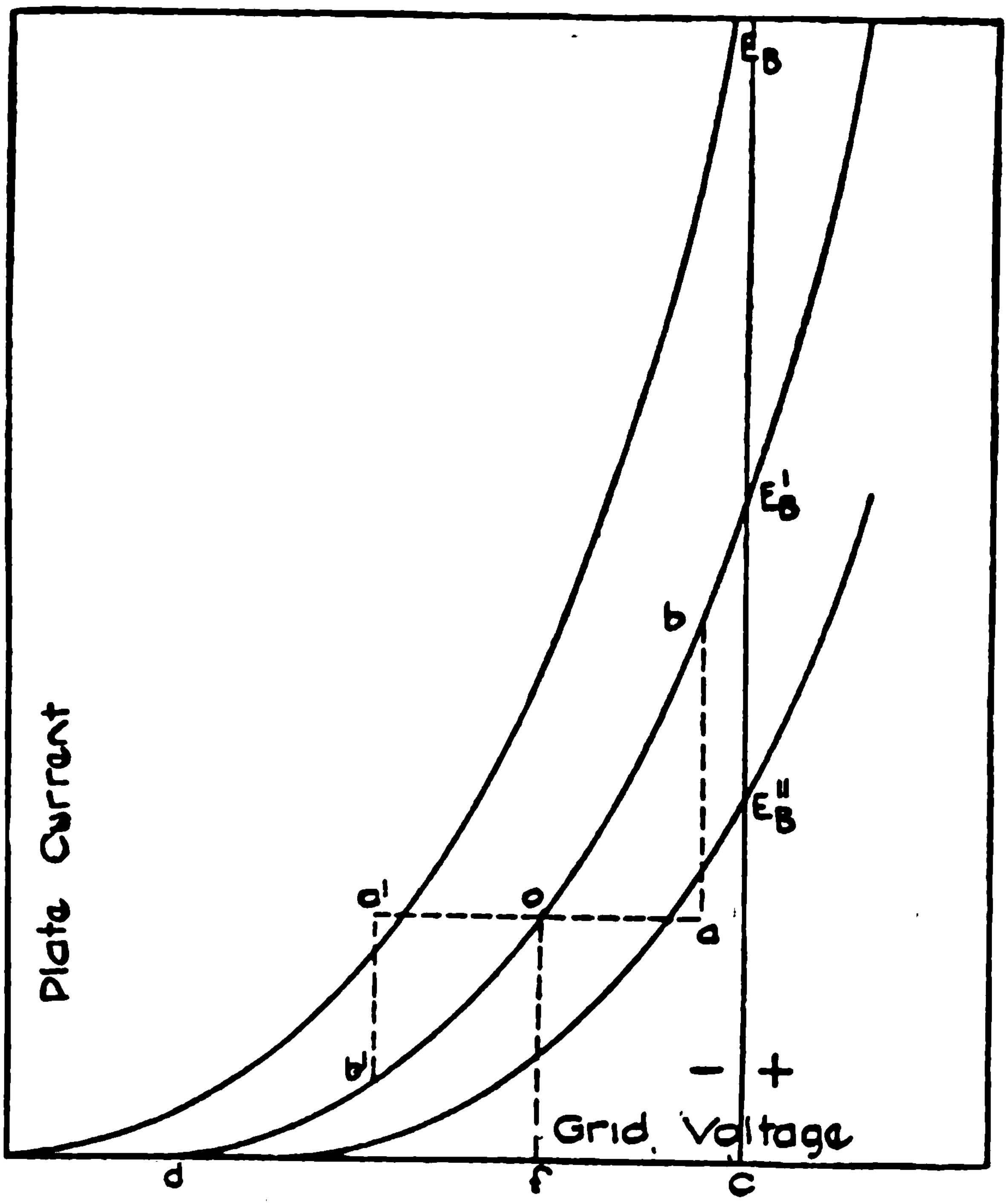

FIGURE 5

Referring to equation (6) and Figure 5, we see that the current is finite for negative values of the grid voltage E_c, and is only reduced to zero when

$$E_c = -(\gamma E_B + \varepsilon).$$

Differentiating I (equation 6), first with respect to E_B, keeping E_c constant, and then with respect to E_c, keeping E_B constant, we get:

$$\frac{\partial I}{\partial E_B} = 2\,\alpha\gamma\,(\gamma E_B + E_c + \varepsilon) = Q, \tag{9}$$

$$\frac{\partial I}{\partial E_c} = 2\,\alpha\,(\gamma E_B + E_c + \varepsilon) = S. \tag{10}$$

Hence

$$\frac{Q}{S}=\gamma=\text{constant}, \tag{11}$$

from which it follows that for equivalent values of E_B and E_c, a change in the anode voltage E_B produces γ times as great a change in the current to the anode as an equal change in the grid voltage E_c.

The output impedance of the tube is obtained from the admittance K which is given by

$$K=\frac{1}{2\pi}\int_0^{2\pi}\frac{\partial I}{\partial E_B}\,dt,$$

or putting in the value of $\frac{\partial I}{\partial E_B}$ from (9):

$$K=\frac{1}{2\pi}\int_0^{2\pi}2\,a\,\gamma\,(\gamma E_B+E_c+\varepsilon+e\sin p\,t)\,d\,t.$$

It is seen that $\frac{\partial I}{\partial E_B}$ is not constant but depends upon the instantaneous value of the input voltage $e\sin p\,t$. This is also obvious since the characteristic is curved. The admittance and impedance, however, are independent of the input voltage, as is seen readily by integrating the expression for K:

$$R_o=\frac{1}{K}=\frac{1}{2\,a\,\gamma\,(\gamma E_B+E_c+\varepsilon)}. \tag{12}$$

Comparing this with equation (9), it is seen that the impedance can readily be obtained by taking the slope of the characteristic at a point corresponding to the direct current values E_B and E_c at which it is desired to operate the tube.

Equation (12) can be expressed in a more convenient form by multiplying its numerator and denominator by $(\gamma\,E_B+E_c+\varepsilon)$:

$$R_o=\frac{\gamma E_B+E_c+\varepsilon}{2\,a\,\gamma\,(\gamma E_B+E_c+\varepsilon)^2},$$

which, with the help of equation (6) becomes

$$R_o=\frac{E_B+\mu_o(E_c+\varepsilon)}{2I}, \tag{13}$$

where

$$\mu_o=\frac{1}{\gamma}. \tag{14}$$

We shall see that μ_o is the maximum voltage amplification obtainable from the device.

Comparing (12) with (9) it is seen that $R_o = 1/Q$ and therefore from (11) and (16) the slope of the I, E_c-curve is given by

$$S = \frac{\mu_o}{R_o}. \tag{15}$$

This constant is very important. It will be shown later that the quality of the device is determined by the value of S, that is, the slope of the curve giving the current to the plate as a function of the grid voltage.

V. Experimental Determination of the Constants of the Tube and Verification of the Characteristic Equation

In order experimentally to verify equation (6) it is necessary to know the value of the constants γ and ε. Both these constants can be determined by methods which do not depend on the exponent of the equation. The linear stray field relation

$$E_s = \gamma E_B + \varepsilon, \tag{3}$$

which is involved in equation (6) is also independent of the exponent. The constants γ and ε can be determined and the relation (3) tested as follows:

Let us assume an arbitrary exponent β for equation (6):

$$I = a\,(\gamma E_B + E_c + \varepsilon)^{\beta}. \tag{16}$$

Takine the general case in which both E_B and E_c are variable, we have:

$$\frac{dI}{dE_c} = \frac{\partial I}{\partial E_B}\frac{dE_B}{dE_c} + \frac{\partial I}{\partial E_c}.$$

Now

$$\frac{\partial I}{\partial E_B} = a\,\beta\,\gamma\,(\gamma E_B + E_c + \varepsilon)^{\beta-1},$$

$$\frac{\partial I}{\partial E_c} = a\,\beta\,(\gamma E_B + E_c + \varepsilon)^{\beta-1}.$$

Hence

$$\frac{dI}{dE_c} = a\,\beta\,(\gamma E_B + E_c + \varepsilon)^{\beta-1}\left(\gamma\frac{dE_B}{dE_c} + 1\right). \tag{17}$$

Now, let I be constant, then

$$\gamma E_B + E_c + \varepsilon = 0,$$

that is,

$$-E_c = \gamma E_B + \varepsilon = E_s, \tag{18}$$

or
$$\frac{dE_B}{dE_c} = -\frac{1}{\gamma}.$$

Integrating and putting $\frac{1}{\gamma}$ equal to μ_o, we get

$$E_B' = E_B + \mu_o E_c. \tag{19}$$

Equations (18) and (19) are, therefore, independent of the exponent of (6). Equation (18) gives the case in which the current has the constant value zero. It shows that E_s in equation (3) is simply the absolute value of the grid potential E_c, which suffices to reduce the current to the anode to zero when the anode has a potential E_B. (The potentials are referred to that of the filament which is supposed to be grounded.) Referring to Figure 5, we see that equation (18) gives the relation between the intercepts of the curves on the axis of grid potential E_c and the corresponding values of anode potentials E_B. This is the method which I used several years ago to test the linear stray field relation (3). The accuracy with which this relation is obeyed is seen from Figures 3 and 5 of my above mentioned publication.[9]

The factor μ_o, which plays a very important part in the theory of operation of the thermionic amplifier, can be obtained by taking the slope of the curve giving the relation between E_B and E_c in accordance with equation (18). It can be more easily determined with the help of equation (19), which gives the relation between the anode and grid potentials necessary to maintain the current at some convenient constant value. Figure 6 gives results obtained in this manner. The linear relation obtained between E_B and E_c verifies equation (19).

Another method of determining μ_o is with the help of equation (11):

$$\frac{Q}{S} = \gamma = \frac{1}{\mu_o}. \tag{11}$$

S is the slope of the curve giving the current to the anode as a function of the grid potential, and Q the slope of the curve which gives the current as a function of the anode potential. Since both these slopes depend upon the anode and grid potentials E_B and E_c, they must be measured for the same values of E_B and E_c. This method gives quite reliable results but is not as convenient as the one explained above.

[9] "Vehr. d. D. Phys. Ges.," above. For further experimental verification of this relation when applied to the case in which the cathode consists of a hot filament, see H. J. van der Bijl, "Phys. Rev.", (2), 12, 184, 1918.

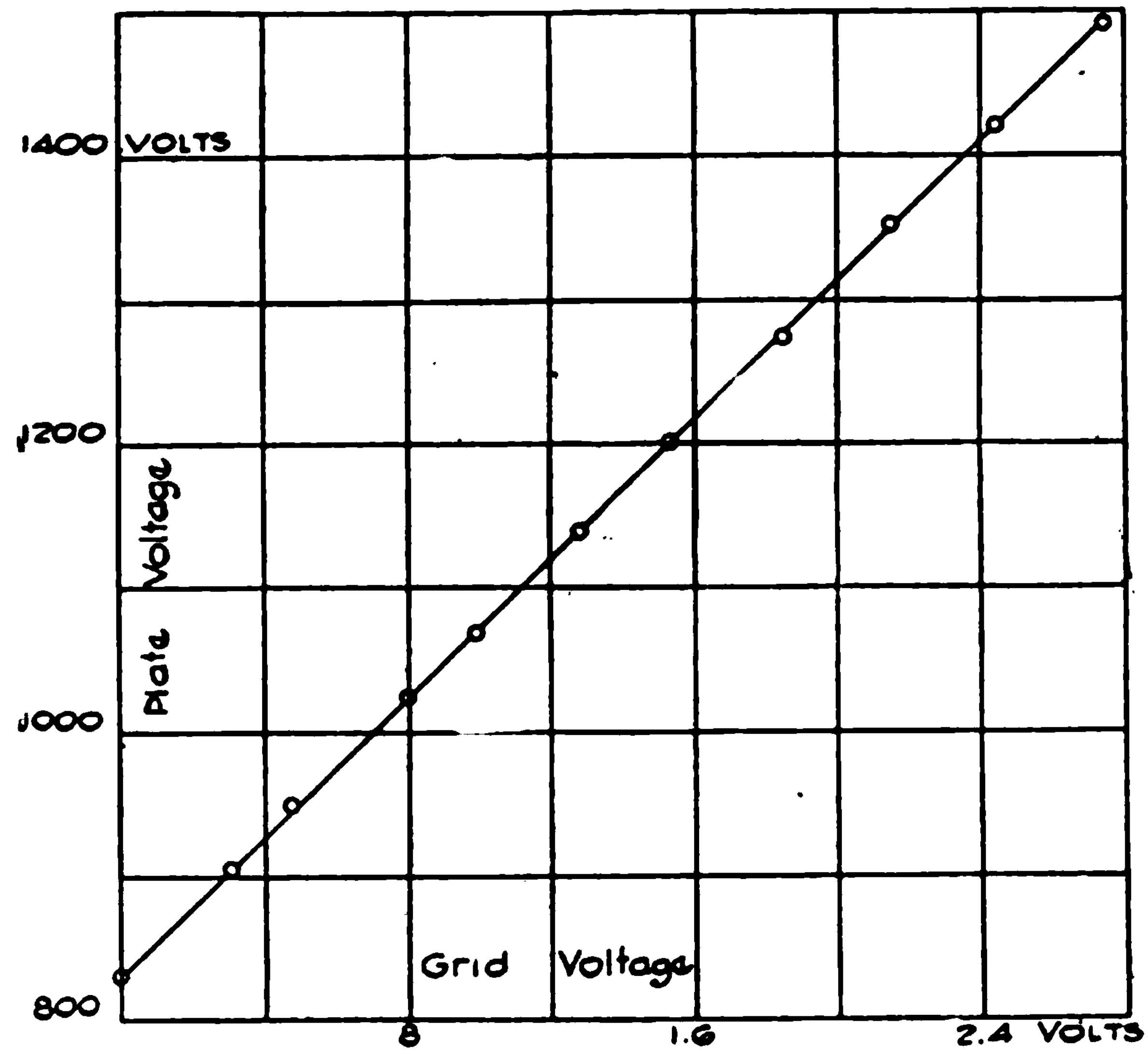

FIGURE 6

The most convenient method of measuring the amplification constant μ_o is that recently given by Miller.[10] The principle of this method is the same as one that has been frequently used in this laboratory where it is necessary to determine μ_o for a large number of tubes. The circuit shown in Figure 7 is contained in a box with terminals for the ammeter A and batteries E_B and E_A. The tube is plugged into a socket provided for it in the box. It is seen from the previous paragraph that a voltage in the grid circuit is equivalent to μ_o-times that voltage in the plate circuit. Hence, referring to Figure 7, it is evident that no change will be produced in the ammeter A on closing the key K, if $\frac{r_1}{r_2} = \mu_o$. For convenience in measurement r_2 is a fixed value of 10 ohms, and r_1 consists of three dial rheostats of 1,000, 100, and 10 ohms arranged in steps of 100, 10, and 1 ohms each. The rheostats are marked in tenths of the actual resistances, so that the setting of the dials read the μ_o directly. The drain of the battery E_1 is very small because the circuit is only closed momentarily by the push button K. This battery, therefore, consists of small dry cells inclosed in the box. Instead

[10] J. M. Miller, PROCEEDING OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 141, 1918.

of using a direct current supplied by the battery E_1, an alternating current can be used, in which case the ammeter A must be replaced by a telephone receiver. The use of an alternating

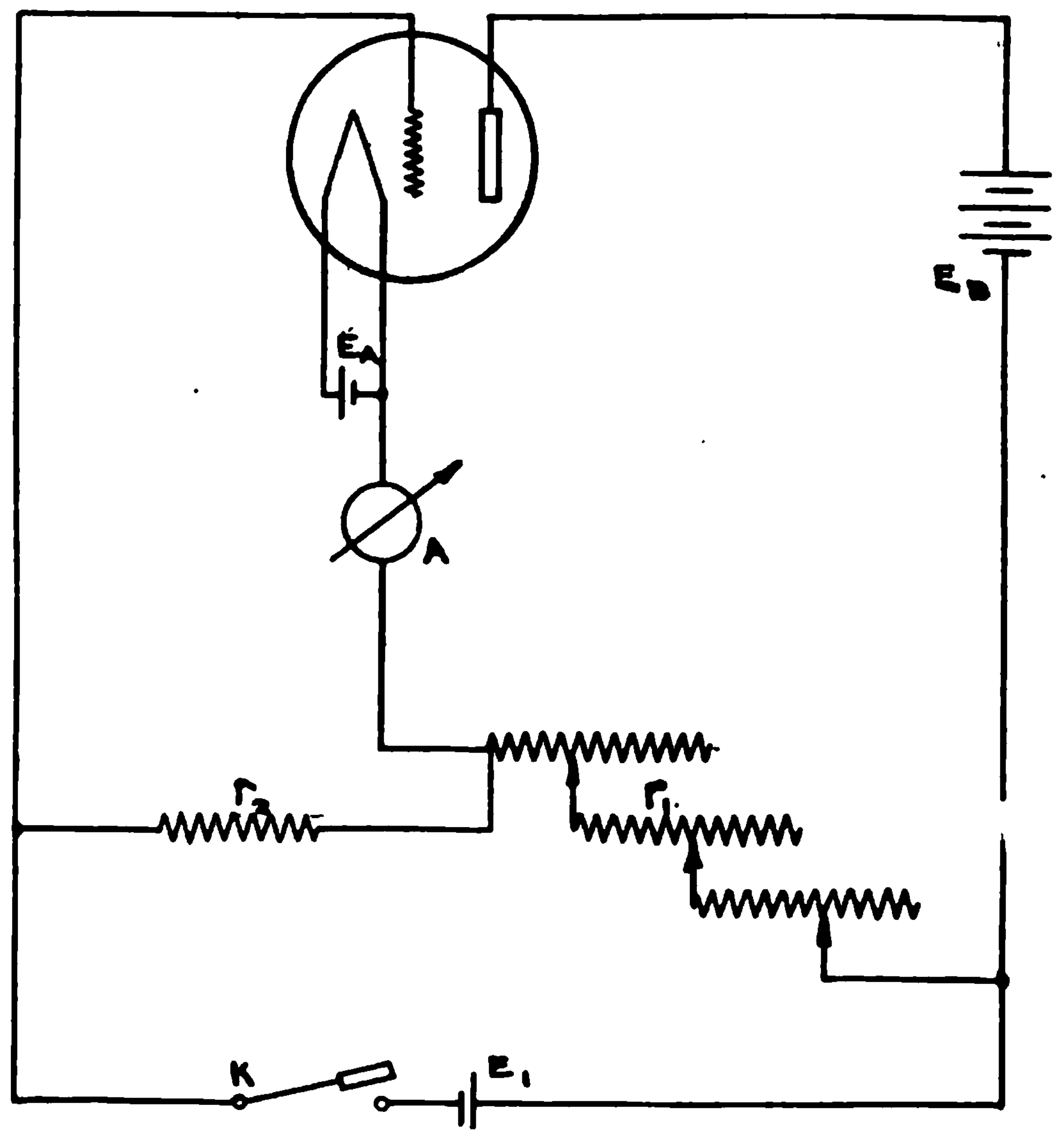

FIGURE 7

current has the advantage that it allows a simple determination of the impedance of the tube according to the method given by Miller. Figure 8 shows a photograph of the μ_0-meter with a tube inserted in its socket. The rheostat R enables the filament current to be adjusted to the desired value.

In order to test the characteristic equation (6) it is still necessary to know the value of ε. This can be obtained by applying a convenient negative potential to the grid and keeping it constant while observing the current to the anode for various values of the anode potential. The grid being negative with respect to the filament, no current could be established in the filament-grid circuit. There should be no resistance in the circuit FPE except that of the ammeter, and this should be small compared with the internal output resistance of the amplifier itself. Under these conditions the voltage of the battery E is always equal to E_B, the voltage between filament and anode, so that the observed values of current and voltage give the true characteristic of the amplifier. From the curve giving the cur-

Figure 8

rent as a function of the anode potential the value of the anode potential can be determined for which the current is reduced to zero. This potential, of course, depends upon that of the grid. By putting I equal to zero in equation (6), and $\gamma = \frac{1}{\mu_o}$ we get

$$\varepsilon = -\left(\frac{E_B}{\mu_o} + E_c\right).$$

Once μ_o and ε are known, the current can be plotted against the expression

$$\left(\frac{E_B}{\mu_o} + E_c + \varepsilon\right)^2 \tag{20}$$

Figure 9 shows the results for one particular type of tube.

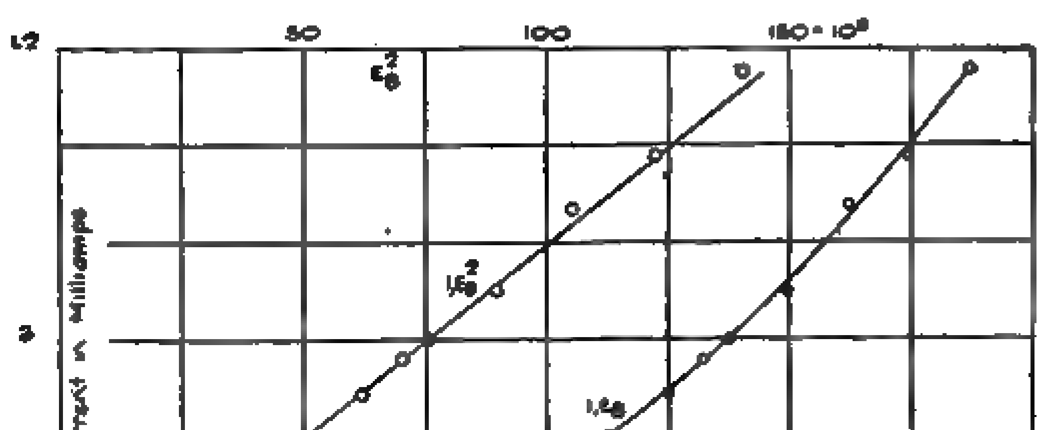

Figure 9

In this case the grid had a constant negative potential equal to ε. Hence the current was simply plotted as a function of E_B. The straight line gives the relation between E_B^2 and the anode current. It is seen that the parabolic relation of the fundamental equation (6) is obeyed with sufficient accuracy.[11]

VI. Characteristic of Circuit Containing Thermionic Amplifier and Ohmic Resistance in Series

In discussing the behavior of the thermionic amplifier in an alternating current circuit, we shall make two assumptions:

First. The alternating current established in the circuit *FPER* (Figure 4) is a linear function of the voltage impressed upon the input circuit $F G E_c$. This implies that the power amplification is independent of the input. This is the condition for an ideal amplifier.

Second. The thermionic amplifier shows no reactance effect. This implies that if the amplifier be inserted in a non-inductive circuit, the power amplification produced is independent of the frequency.

No proof is needed to establish the validity of the second assumption. The first is, however, not true except under certain conditions, and it remains to determine these conditions and operate the amplifier so that they are satisfied. Let the external resistance of the output circuit *FPE* (Figure 4) be zero, and the resistance of the current-measuring device negligibly small compared with the internal output impedance of the amplifier itself. Under these circumstances E_B, which denotes the potential difference between filament and anode, is independent of the current in the circuit *FPE*, and always equal to the voltage E of the battery in the circuit. Hence, if the current be plotted as a function of E_c, the potential difference between filament and grid, the parabola given by equation (6) and Figure 5 is obtained. If, now, the potential of the grid be varied about the value E_c equal to cf (Figure 5), it is obvious from the curve that the increase, ab, in current to the anode due to a decrease, oa, in the negative potential of the grid is greater than the decrease $a'b'$ in current caused by an equal increase, oa', in the negative grid potential. In this case, the output current consists of the following parts: Let the alternating input voltage superimposed upon E_c be $e \sin p t$ then

$$I_o = a\,(\gamma E_B + E_c + \varepsilon + e \sin p t)^2. \qquad (8)$$

[11] For further results of these experiments, see "Phys. Rev.", (2), 12, 186, 1918.

Expanding this we get

$$I_o = \alpha(\gamma E_B + E_c + \varepsilon)^2 + 2\alpha(\gamma E_B + E_c + \varepsilon)\, e \sin pt$$
$$+ \frac{\alpha e^2}{2} \cos(2pt + \pi) + \frac{\alpha e^2}{2}. \tag{21}$$

The first term represents the steady direct current maintained by the constant voltages E_B and E_c when the input voltage e is zero (equation 6). The second term gives the alternating output current oscillating about the value of direct current given by (6). It is in phase with and has the same frequency as the input voltage. When using the device as an amplifier, this is the only useful current we need to consider. The first harmonic represented by the third term is present, as was to be expected in virtue of the parabolic characteristic. The last term, which is proportional to the square of the input voltage, represents the change in the direct current component due to the alternating input voltage, and is the only effective current when using the device as radio wave detector. If a direct current meter were inserted in the output circuit, it would show a current which is greater than that given by equation (6) by an amount equal to $\frac{\alpha e^2}{2}$, the last term of equation (21). This is the state of matters when the device works into a negligibly small resistance.

If, on the other hand, the output circuit contains an appreciable resistance,[12] R, the voltage E_B between filament and plate is not constant, but is a function of the current, and an increase in the current due to an increase in the grid potential sets up a potential drop in the resistance R, with the result that E_B decreases, since the battery voltage E is constant. E_B is now given by

$$E_B = E - RI. \tag{22}$$

In order to obtain the characteristic of the circuit containing the tube and a resistance R, let us substitute (22) in (8):

$$I_o = \alpha[\gamma(E - RI) + E_c + \varepsilon + e \sin pt]^2$$

and put $\gamma E + E_c + \varepsilon = V$. This gives

$$I_o = \frac{1 + 2\alpha\gamma R(V + e \sin pt) - \sqrt{1 + 4\alpha\gamma R(V + e \sin pt)}}{2\alpha\gamma^2 R^2}. \tag{23}$$

[12] The insertion of a suitable resistance in the output circuit to straighten out the characteristic and so reduce distortion was, I believe, first suggested by Dr. Arnold, who also showed experimentally that distortion is almost negligible when the external resistance is equal to the impedance of the tube.

This expression can be expanded into a Fourier series:

$$I_o = \left\{ \begin{array}{l} \left(\dfrac{1+2\,\alpha\gamma R V - \sqrt{1+4\,\alpha\gamma R V}}{2\,\alpha\gamma^2 R^2} \right) \\ + \left(1 - \dfrac{1}{\sqrt{1+4\,\alpha\gamma R V}} \right) \dfrac{e}{\gamma R} \sin p t \\ + \dfrac{(-1)^{n+1}\, 2^n\, (2n-1)!\, \alpha^n \gamma^{n-1} R^{n-1} e^{n+1} \sin^{n+1} p t}{(n+1)!\, (1+4\,\alpha\gamma R V)^{2n+\frac{1}{2}}} . \end{array} \right. \tag{24}$$

From this it is seen that the rate of convergence of the series increases as R is increased. Actual computations show that when the tube is made to work into an impedance equal to or greater than that of the tube the harmonic terms become negligibly small compared with the second term of (24), which is the only useful term when using the tube as an amplifier, so that we can assume that the amplification is independent of the input voltage.[13]

When the tube works into a large external resistance it can show a blocking or choking effect on the current. This is seen from the following: The voltage E_B which is effective in drawing electrons thru the grid to the anode is given by equation (22). If now the current be increased, not by increasing the electromotive force in the circuit *FPR* (Figure 5), but by increasing the potential difference between filament and grid, the current I increases, while E, the electromotive force in the plate circuit, remains constant, from which it follows that E_B must decrease while E_c, the grid voltage, increases. The result of this is that more electrons that otherwise would have come thru the grid to the anode are now drawn to the grid. If the input voltage becomes large enough the anode circuit *FPR* may be robbed of so many of its electrons that no further increase in current in the anode circuit results no matter how much the grid voltage is increased. Under these conditions equation (24) does not apply. Its application is limited to the conditions stated by equations (25) and (26).

Even if the series represented by the last term of equation

[13] In this connection I want to point out that altho the parabolic relation used here represents the characteristic of the tube with sufficient accuracy when using the tube as an amplifier, and indeed with quite a good degree of accuracy, as shown by the experimental curves, yet the approximation is not close enough to represent accurately the second and higher derivatives of the characteristic, and therefore, too much reliance should not be placed on the actual values of the several harmonic terms represented by the last term of equation (24). This equation is merely intended to show, as it does, in a general way how the insertion of a resistance in the output circuit of the tube tends to straighten out the characteristic.

(24) were zero, distortionless transmission can only be obtained if the input voltage is kept within certain limits.

Let the input voltage, $e \sin pt$, be superimposed upon the negative grid voltage, E_c (Figure 5). Theoretically speaking, one condition of operation is that the grid should never become so much positive with respect to the filament that it takes appreciable current, for if this happens the current established in the grid circuit would lower the input voltage, and therefore the amplification. In actual practice the extent to which the grid can become positive before taking appreciable current depends upon the value of the plate voltage and the structure of the tube. We can therefore state that a condition for distortionless transmission is $e \leqq |E_c| + |g|$, where g is the positive voltage which the grid can acquire without taking enough current to cause distortion. Another condition is that the input voltage must not exceed the value given by df (Figure 5); otherwise the negative peaks of the output current wave will be chopped off. Now cd is given by $\gamma E_B + \varepsilon$. This is obtained by equating the current I to zero in equation (6). We therefore have the conditions

$$\left.\begin{aligned} e &\leqq |E_c| + |g|, \\ e &\leqq |\gamma E_B + \varepsilon| - |E_c|, \end{aligned}\right\} \tag{25}$$

or when the tube is working at full capacity—that is, when operating over the whole curve,

$$e = |E_c| + |g| = |\gamma E_B + \varepsilon| - |E_c|. \tag{26}$$

It may be remarked here that when using the tube as an oscillation generator, these limits are not obeyed. From equation (12), it is seen that the impedance R_o of the tube is independent of the input voltage $e \sin pt$. This is, however, true only as long as the characteristic is parabolic. Referring to Figure 5, if the input voltage oscillates about the value f, the slope of the tangent at o is a measure of the impedance, in fact, it is μ_o divided by the impedance (equation 15). Since the characteristic is parabolic, it follows that the secant thru $b\,b'$ is always parallel to the tangent at o as long as $oa = oa'$ (equal to the input voltage e). The impedance can, therefore, be obtained by taking the slope of the secant thru the maximum and minimum current values. If now the tube works beyond the limits of the parabolic characteristic, such as along the curve OAB (Figure 2) the slope of this secant does not remain constant for all values of the input voltage, so that the impedance of the tube is not independent of the strength of the oscillations but increases with it, the minimum impedance being obtained when the oscilla-

tions are infinitely small. This is what happens when the tube operates as an oscillation generator. Part of the energy in the output circuit is fed back to the input circuit, thus increasing the strength of the oscillations in the output until the tube works beyond the limits of the parabolic characteristic. The impedance of the tube is thereby increased, and this increase continues until the impedance acquires the maximum value capable of sustaining the oscillations, consistent with the degree of coupling used between the output and input circuits. It is readily seen that the current obtained in the output is not a single pure sine wave, but contains a number of harmonics as well. The extent to which these harmonics influence the current values obtained, when the latter is measured simply by the insertion of a hot wire meter in the output circuit, which, of course, measures the total current, depends on the degree of coupling as well as the constants of the output circuit. These considerations must be borne in mind when dealing with the alternating current output power obtainable from an oscillation tube. The only useful power is, of course, that which is due to the fundamental.

VII. Amplification Equations of the Thermionic Amplifier

On the strength of the two assumptions discussed in the previous paragraph, namely, that the amplification is independent of the input and the frequency, it is possible to derive the equations of amplification in a very simple way. Referring to Figure 4, let the current in the external resistance R be varied by variations produced in the grid potential, E_c. Then, as was shown in the last paragraph, E_B is also a variable depending on the current I, as shown by

$$E_B = E - R\,I, \tag{22}$$

where E is the constant voltage of the battery in the output circuit $FPER$. Hence

$$I = \Phi\,(E_B, E_c),$$

from which

$$\frac{dI}{dE_c} = \frac{\partial I}{\partial E_B} \cdot \frac{dE_B}{dE_c} + \frac{\partial 1}{\partial E_c}.$$

This gives the variation of current in R as a function of the variation in the grid voltage.

Substituting from (9) and (10),

$$\frac{dI}{dE_c} = 2\,a\,(\gamma E_B + E_c + \varepsilon)\left(\gamma\,\frac{d(E-RI)}{dE_c} + 1\right),$$

that is,

$$\frac{dI}{dE_c} = \frac{2\,a\,(\gamma E_B + E_c + \varepsilon)}{1 + 2\,a\,\gamma\,R\,(\gamma E + E_c + \varepsilon)}.$$

Multiplying thruout by R, and putting $\gamma = \frac{1}{\mu_o}$ we obtain by a simple transformation

$$R\frac{dI}{dE_c} = \frac{\mu_o R}{R + \frac{E_B + \mu_o(E_c + \varepsilon)}{2I}}. \tag{27}$$

Now, $R \cdot dI$ is the voltage change set up in the resistance R, and dE_c is the change in the input voltage. Hence equation (27) gives the voltage amplification produced by the device, which we shall call μ. Furthermore, it follows from paragraph IV. that the output impedance of the amplifier is given by

$$R_o = \frac{E_B + \mu_o\,(E_c + \varepsilon)}{2I}. \tag{13}$$

Hence the voltage amplification μ is given by

$$\mu = \frac{\mu_o R}{R + R_o}. \tag{28}$$

From this equation it is seen that the voltage amplification asymptotically approaches a finite value μ_o, which is attained when the external resistance R becomes infinitely large compared with the output impedance of the amplifier.

In order to find the power amplification it is necessary to know the input impedance of the amplifier; that is, the impedance of the circuit $F\,G\,E_c$ (Figure 4). Now, the amplifier is operated, as was stated above, under such conditions that no current is established in the circuit $F\,G\,E_c$. The impedance of this circuit is, therefore, infinite, and the power developed in it is indeterminate.

In order to give the input circuit a definite constant resistance, Mr. Arnold suggested shunting the filament and grid with a high resistance. This can be considered as the input resistance R_i, of the amplifier. The input voltage is that developed between the ends of this shunt resistance.

If, now, e and e_i represent the voltages established between the ends of the output and input resistances R and R_i respectively, the power developed in R and R_i is $\frac{e^2}{R}$ and $\frac{e_i^2}{R_i}$. Hence the power amplification is

$$\eta = \frac{e^2 R_i}{e_i^2 R} = \mu^2 \frac{R_i}{R},$$

which, with the help of (28), becomes

$$\eta = \frac{\mu_o^2 R_i R}{(R+R_o)^2}, \tag{29}$$

The amplification is, therefore, a maximum when R is equal to R_o.[14]

The power developed in R is

$$P = \frac{\mu_o^2 e_i^2 R}{(R+R_o)^2}, \tag{30}$$

from which it follows, as was to be expected, that the power in R is a maximum when the external output resistance R is equal to the output impedance R_o of the tube.

It is readily seen that the current amplification is given by

$$\xi = \frac{\mu_o R_i}{R+R_o}, \tag{31}$$

from which it follows that the current amplification asymptotically approaches zero as R is increased, the maximum current amplification being obtained when R becomes infinitely small compared with R_o.

Putting $R=R_o$ in (29) and $R=0$ in (31) and remembering that the slope of the curve giving the relation between plate current and grid voltage is given by

$$S = \frac{\mu_o}{R_o}, \tag{15}$$

we get for the maximum power amplification

$$\eta' = \frac{\mu_o R_i}{4} \cdot S, \tag{29a}$$

and for the maximum current amplification

$$\xi' = R_i \cdot S. \tag{31a}$$

These equations show the important part played by the slope S of the curve giving the plate current as a function of the grid voltage. The factor S is equally important in the operation of tube as an oscillation generator and detector. The slope S is what was called by Hazeltine the "mutual conductance" of the tube.[15]

[14] The amplification equations are derived on the assumption that the external output circuit of the tube contains only pure resistance. When the circuit is reactive, as is common in practice, we can, to a first approximation, substitute the effective impedance for R in the equations. Mathematical proof for this is given by J. R. Carson, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, number 2, April, 1919.

[15] L. A. Hazeltine, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 63, 1918.

It is seen here that this quantity can be expressed in terms of the two most important and easily determined constants of the tube, namely, the amplification constant μ_o and the impedance R_o.

While the amplification constant depends only upon the structure of the tube, the impedance, besides being a function of the structure, depends also upon the values of the applied voltages between filament and grid and filament and plate, as is readily seen from the above equation. If the impedance is determined as a function of the plate voltage, the grid voltage being, let us say, zero, a curve is obtained somewhat like that shown in Figure 10. If now it is desired to operate the tube

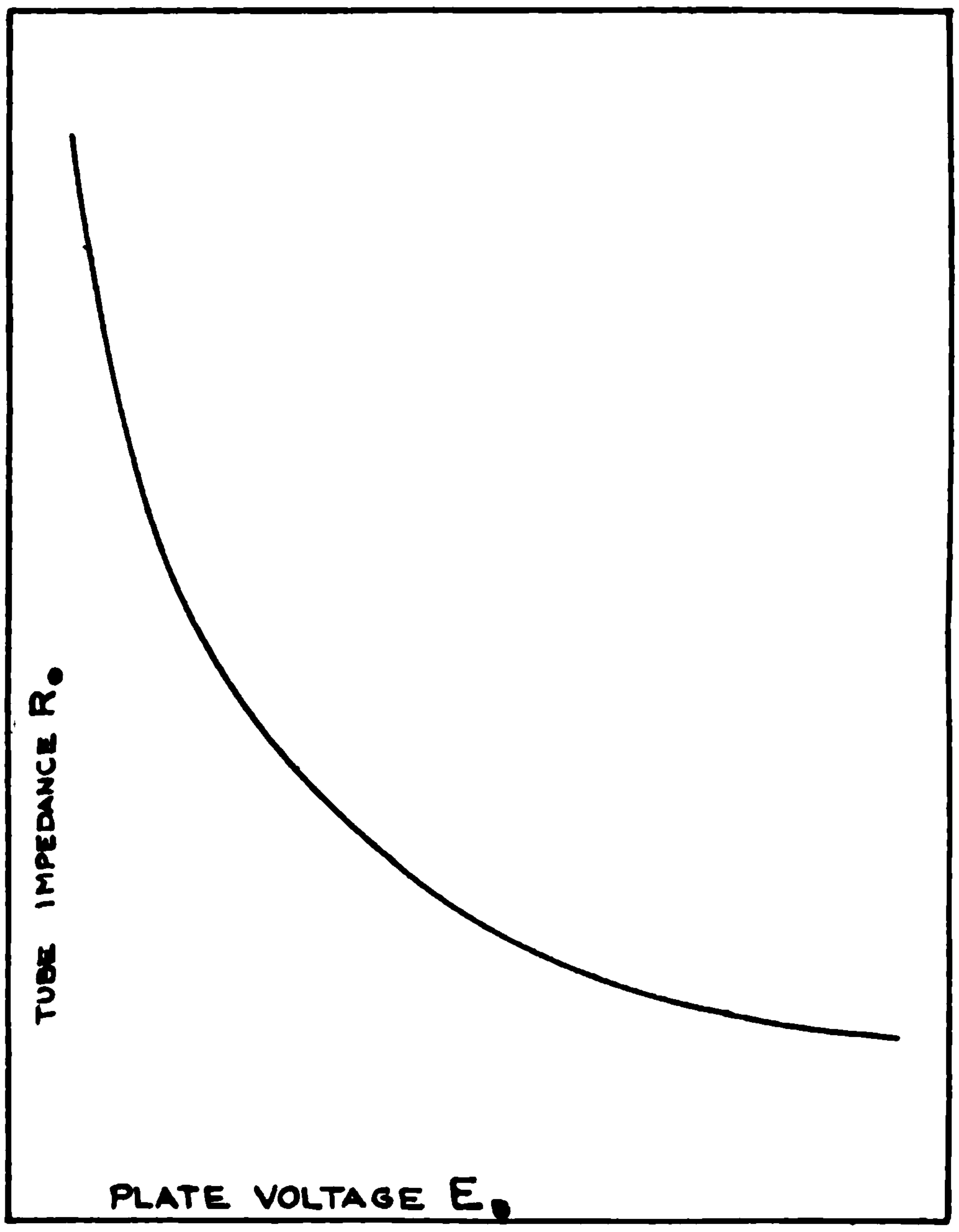

FIGURE 10

with a definite plate voltage E_B and a grid voltage E_c other than zero, the impedance under these conditions can be obtained from such a curve by adding $\pm\mu_o E_c$ to E_B and reading off the impedance from the curve at a value of the plate voltage equal to $E_B \pm \mu_o E_c$.

In designing a tube, the structural parameters are so chosen

that the tube constants have definite values depending upon the purpose for which the tube is to be used. Suppose it is desired to design a two-stage amplifier set. Since the tube is a potential-operated device, the voltage impressed on the input of the tube must be made as high as possible, irrespective of the value of the current in the input circuit. This is usually done by stepping up the incoming voltage by means of a transformer, the secondary of which is wound to have as high an impedance as possible. For the same reason when the output current of one tube is to be amplified by another, the first tube is made to work into an impedance or resistance which is large compared with its internal output impedance. Such an arrangement allows of a large voltage amplification being obtained from the first tube. This follows from equation (28) which also shows that when used as a voltage amplifier, the tube must have a large amplification constant μ_o. Referring to equation (26), however, it is seen that the larger μ_o is, the smaller is the input voltage e that can be impressed on the tube without producing distortion, provided that E_B is fixed. When it is necessary to use a two- or three-stage amplifier set, the incoming voltage is generally so small that μ_o for the first tube can be quite large and the plate voltage still not excessively high. But then the voltage impressed on the second tube is much larger than that impressed on the first, and the second tube must be so designed as to be capable of handling this voltage. If the first tube, for example, has an amplification constant equal to 40 and works into a resistance four times its own impedance, the voltage on the second tube is 32 times that impressed on the first. It is seen, therefore, that unless the plate voltage on the second tube be made very much higher than that on the first, the two tubes must have entirely different structural parameters, if equation (26) is to be satisfied in both cases. Such considerations show that unless the tubes be properly designed and the plate voltages correctly chosen to satisfy equation (26), there is a practical limit to the number of stages of amplification that can be used. If the limitations imposed by equation (26) are not taken regard of, the process of amplification can result in a considerable amount of distortion, which is a serious matter when using the device for amplifying telephonic currents. When using it for telegraph purposes, such as the amplification of radio telegraph signals at the receiving station, the distortion produced results in a waste of energy in harmonics.

VIII. Experimental Verification of Amplification Equations

The circuit shown in Figure 11 is not of the type customarily used in practice, but was designed to test the equations developed in the previous paragraph. This type of circuit was

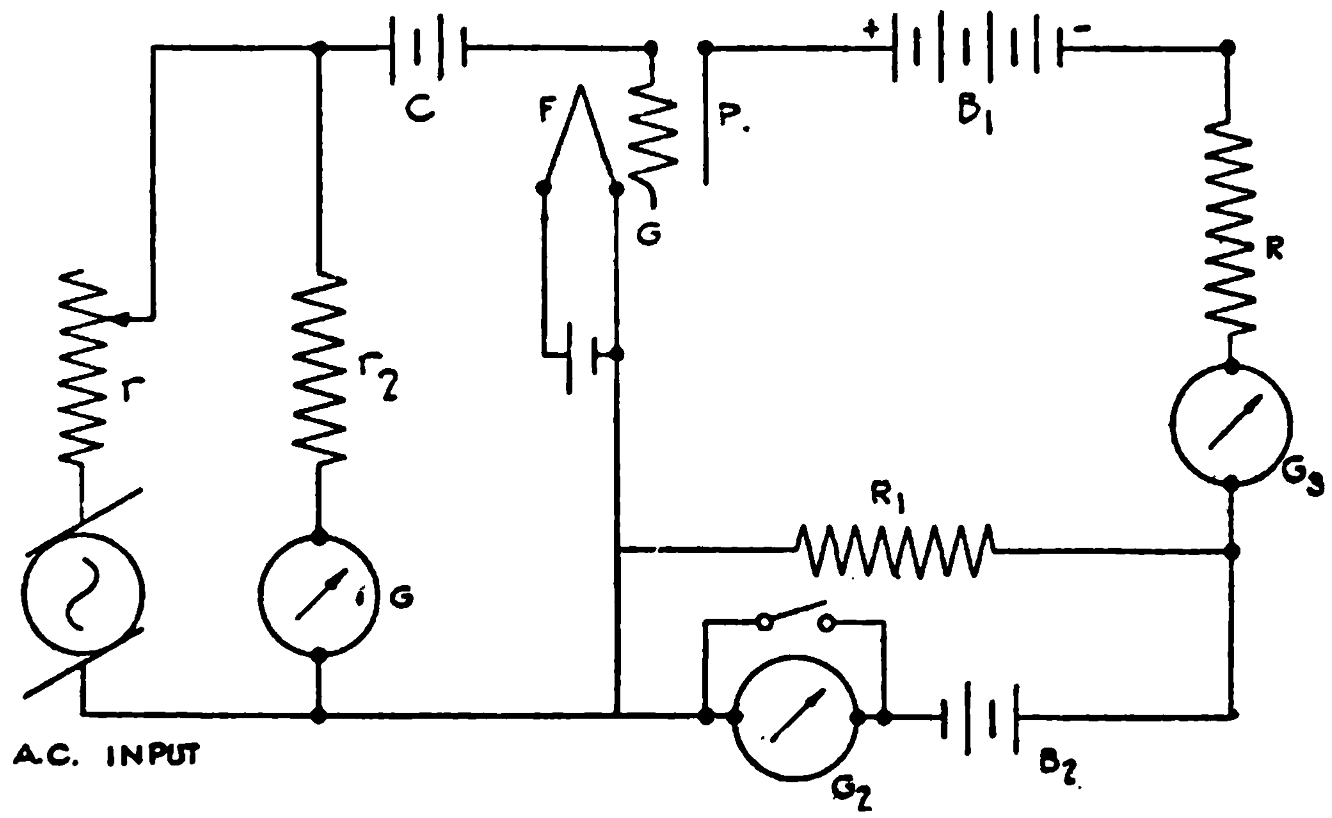

Figure 11

necessitated by the following reasons. Referring to equation (24), it is seen that if the input voltage $e \sin pt$ is zero, the current thru the tube is given by the first term of the equation, which is larger than the alternating current term. For finite values of $e \sin pt$, the resulting alternating current established in the output circuit, which is to be measured, can not be separated in the usual way from this direct current with the help of appropriate inductances and capacities, since then the measured amplification would be largely determined by the constants of the circuit. On the other hand, it is not possible to make the amplifier work simply into a straight non-inductive resistance alone, since the direct current that would flow thru the galvanometer is in most cases large compared with the output alternating current, so that a galvanometer which would be capable of carrying the direct current would not be sensitive enough to measure the output alternating current with any degree of accuracy. This was overcome by using a balancing circuit shown in Figure 11. R and R_1' are two non-inductive resistances stretched upon a board. Parallel to R_1 was shunted a sensitive alternating current galvanometer, G_2, and a balancing

battery B_2. This battery was so adjusted that when no alternating current input was applied to the tube, the current thru the galvanometer was zero, that is, the direct current in the output circuit went thru the resistance R_1. This resistance was large compared with that of the galvanometer; hence practically all the alternating current established in the output when the input voltage was impressed, went thru the galvanometer G_2. The effective resistance into which the tube worked was, of course, given by R. The input was varied with the help of the resistances r_1 and r_2. The whole system was carefully shielded and care was taken to avoid any effects due to shunt and mutual capacity of the leads and resistances. The input voltage was varied from a few hundredths of a volt to several volts and the frequency from 200 to 350,000 cycles per second.

Some of the results are shown in the following figures. Figure 12 shows the output voltage (that is, the voltage across the external resistance R) as a function of the input voltage for a frequency of 1,000 cycles per second. The linear relation indi-

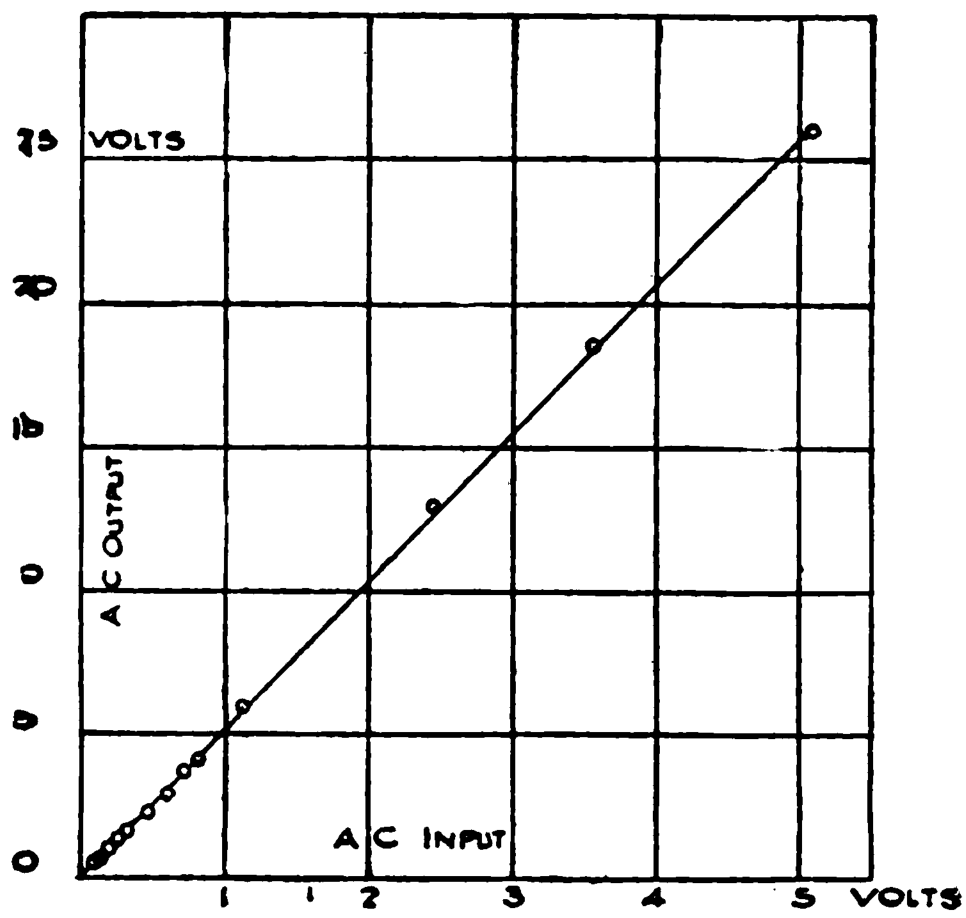

FIGURE 12

cates that the voltage amplification is independent of the input voltage; hence also the power amplification is independent of the input power.

Figure 13 shows the results obtained when the voltage amplification was measured as a function of the external resistance R. The circles show the observed values, while the curves were

calculated from equation (28). It is seen that the agreement is quite good. In this case the input voltage was 0.45 volt and the value of μ_o for this tube, as measured by the direct current method explained in paragraph V. was 10.2.

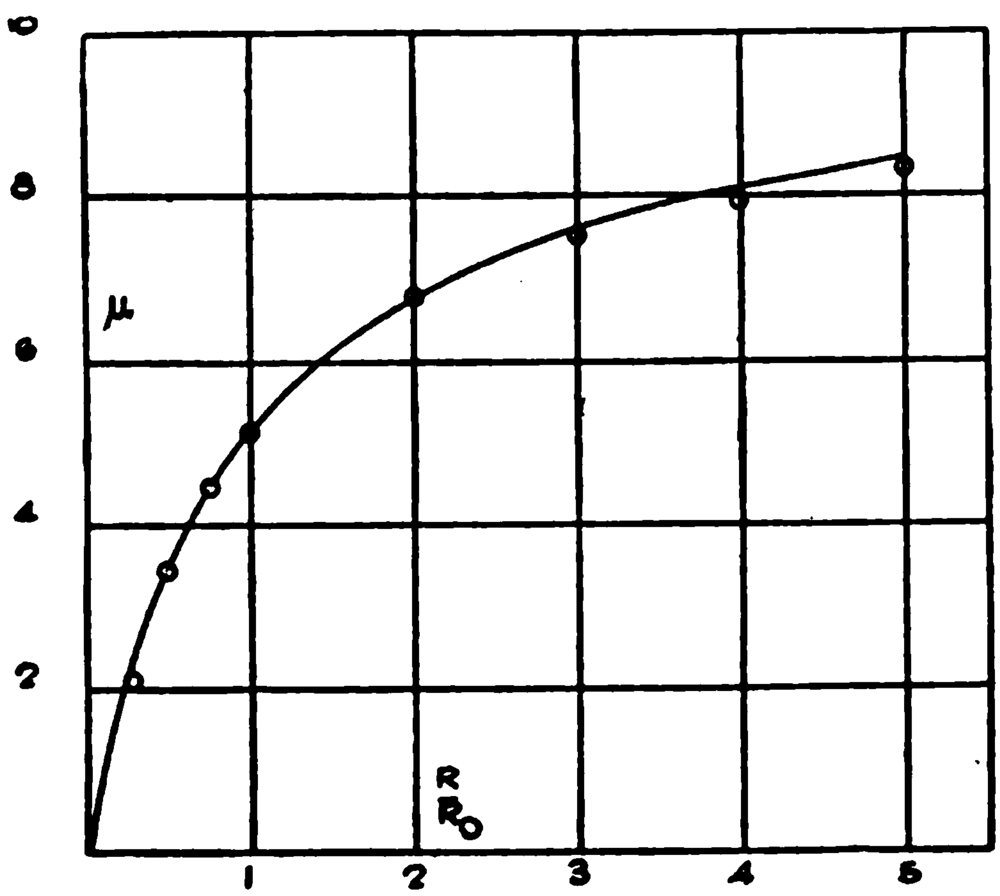

FIGURE 13

$\mu_o = 10.2$, input $= 0.45$ volt

In another experiment, the output power was determined as a function of the external resistance. According to equation (30) this should be a maximum when the external resistance R is equal to the impedance R_o of the tube. In this case the input voltage e_i was 3.55 volts, and $\mu_o = 10.2$. The impedance of the tube was kept constant at 14,800 ohms. This was done by always adjusting the plate voltage so that when the external resistance was changed the current thru the tube was kept constant. From the results given in Figure 14 it is seen that the maximum occurs at $R = 15{,}000$, which is very nearly equal to the impedance of the tube. This result is in accordance with equation (30). Furthermore, the maximum power computed from equation (30) is 22.2 10^{-3} watt, which is sufficiently close to the observed value, 23 10^{-3} watt, to verify equation (30). This equation does not give the maximum power that can be handled by the tube but merely the power developed in the external resistance R for a given value of the input voltage e_i. The maximum power is obtained when the input voltage has the value given by equation (26).

The circuit shown in Figure 11 was used merely to test the equation derived in this paper. It is not suitable for practical purposes where it is necessary to test a large number of tubes with the speed and facility called for in practice. If the tube is to be used as a voltage amplifier all that is necessary is a determination of the amplification constant μ_0. The actual voltage amplification

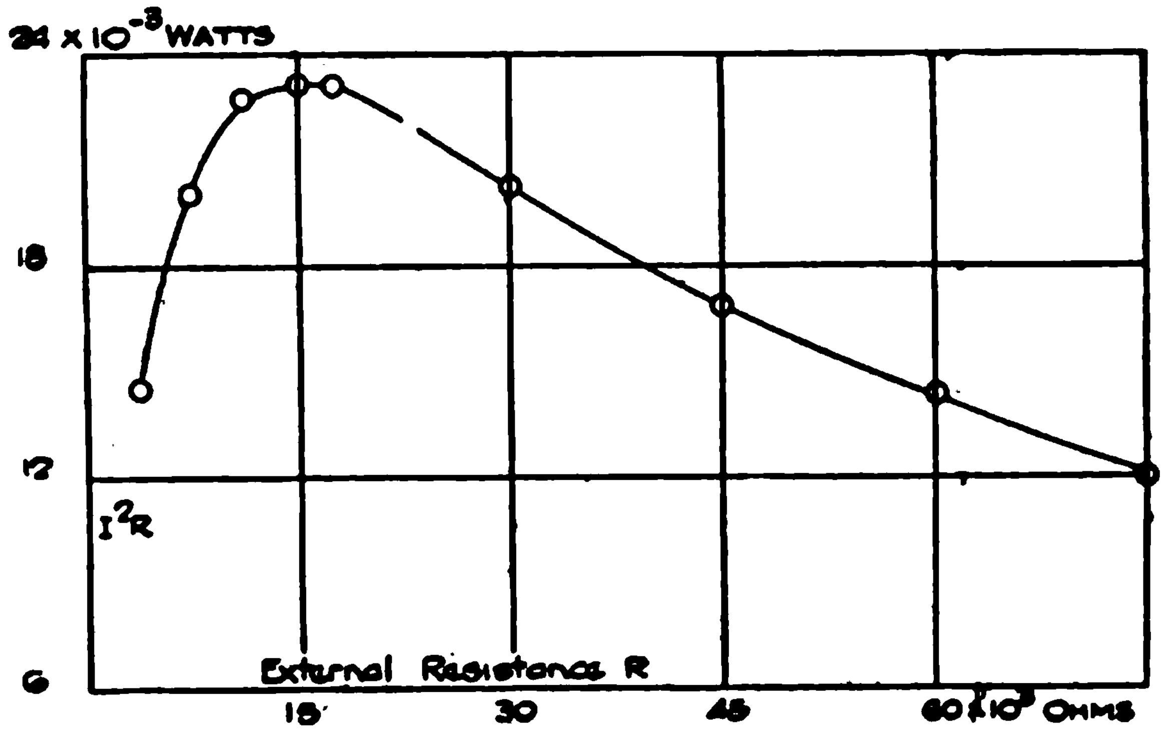

FIGURE 14

obtainable in any particular circuit can then be deduced from equation (28). When the tube is to be used as a power amplifier, a transmission test is used which is common in telephone practice. If it is desired to operate the tube as power amplifier in a certain circuit, its amplification is tested in an equivalent circuit which is arranged so that a note of, say, 800 cycles can be transmitted either straight to a telephone receiver or, by throwing a switch, to the receiver thru the tube and an artificial telephone line, the attenuation of which can be adjusted until the note heard in the receiver is of the same intensity for both positions of the switch. When this is the case the amplification given by the tube is equal to the attenuation produced by the line. The attenuation can be computed from the constants of the line and is usually expressed in terms of miles length of cable of specified constants. This method of measuring and expressing the amplification is convenient in practice. But it must be remembered that this notation has very little meaning and is apt to lead to confusion unless the constants of the cable are definitely specified or previously agreed upon. It simply means that the amplification is equivalent to the attenuation which would be produced by so many miles of a certain sort

of cable having a certain definite attenuation constant. Thus, the current amplification produced by the tube can be expressed in terms of length, d, of cable by the following equation:

$$d = K \log_{10} \frac{i_2}{i_1}.$$

where K is determined by the attenuation of that cable. When dealing with power amplification K must be divided by 2. If we adopt as standard the so-called "standard number 19 gauge cable" used by the Western Electric Company, the length, d, is expressed in miles when the constant, K, has the value $K = 21.13$. The fact that this constant is already finding its way into common vacuum practice would suggest its general adoption when speaking of the amplification of a tube. On the other hand, since the unit of measurement is not a cable but a constant, it might have been more desirable to adopt for K the simple value 20, which would have simplified computations. However, the main point is that it is very important to have a common agreement on the value of K.

The foregoing considerations show that the structural parameters play a very important part in the operation of the tube. On them depend the constants μ_o and R_o which appear in the amplification equations and which are involved explicitly and implicitly in the fundamental equation of the characteristic (equation 6). Proper structural design manifests many latent possibilities of this type of device, and enables us to meet the many conditions that must be complied with in order to obtain satisfactory operation in its ever-increasing number of applications. By proper choice of the structural parameters, tubes have been designed to have voltage and power amplification covering a wide range. A power amplification of 3,000-fold was found possible with a single tube using a plate voltage of only 100 volts. It is not difficult to obtain a voltage amplification of several hundred fold, but in building tubes of such high voltage amplification regard must be taken of the increase in impedance with increase in μ_o, as shown by equations (12) and (13).

Altho the simple theory of operation given in this paper applies specifically to the case in which the tube is used as an amplifier, it has also been of considerable help in designing and developing vacuum tube oscillation generators and detectors. The design of a good detector tube depends very much upon operating conditions. The detecting qualities of a tube can easily be increased by designing it to operate on

comparatively high voltages. This is sometimes desirable in radio stations where high voltages are available and it is desired to use the heterodyne method of reception. On the other hand, it is often important to use tubes of such design that they can operate efficiently on low voltages and with small power consumption. In this connection it may be said that detectors have been designed to give satisfactory operation with two volts on the filament and a plate voltage of 6 volts and less. In the absence of any satisfactory way of expressing the efficiency of a detector, it is unfortunately not possible to say just how good such a detector is. The problem of measuring the detecting efficiency will be reserved for a future paper.

The indebtedness of the writer is due to Mr. E. H. Colpitts and Mr. H. D. Arnold for valuable advice and kind interest which greatly facilitated the work; and to Mr. H. W. Everitt for able assistance in carrying out the experiments.

SUMMARY: The theory of operation of the three-electrode thermionic vacuum tube given in this paper is based on the fundamental equation of the family of characteristic curves for various plate and grid voltages. This equation, which may be written

$$I = a\left(\frac{1}{\mu_o}\Sigma E_B + \Sigma E_c + \varepsilon\right)^2$$

where E_b and E_c are the plate and grid voltages respectively, and a, μ_o and ε are the structural parameters of the device, is obtained empirically with the help of the relation previously discovered by the author, which states that the voltage between filament and plate bears a linear relation to the effective voltage produced by it between the filament and a plane coincident with that of the grid. From this it follows that an electromotive force e impressed upon the grid circuit produces an electromotive force $\mu_o e$ in the plate circuit. With the help of these fundamental relations the amplification equations of the tube are derived in terms of the structural parameters of the tube and the constants of the circuit.

Methods are given for experimentally determining the constants of the tube, the two most important of which are the amplification constant μ_o and the internal output impedance.

Experiments are described which were performed to test these equations, and they indicate that the first order approximation made give results sufficiently accurate for amplification purposes.

THE OPERATIONAL CHARACTERISTICS OF THERMIONIC AMPLIFIERS*

By

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INTRODUCTION

Remarkable strides have been made in the past few years in connection with the definition and measurement of the operational constants of three electrode vacuum tubes. This has given great impetus to the transition of the device from the domain of the scientist to that of the engineer, with the result that it is now possible to design intelligently vacuum tubes for quite a variety of purposes. With the definition of the tube constants well in mind, and with the aid of appropriate methods for measuring them, we may proceed to accumulate by careful research, data connecting the variation of physical dimensions with the constants so defined, which will serve adequately as a basis for later design work. For this purpose, a very interesting course of procedure would consist in holding the plate-filament distance constant and varying the grid-filament space in small steps, plotting the constants against the latter distance as an independent variable with the plate-filament distance taken as a parameter. The information accruing from an investigation of this sort is of the greatest engineering value. Its accumulation has been made possible by the definitions and methods of measurement that have been developed by such workers as Hazeltine,[1] Langmuir,[2] Miller,[3]

* Received by the Editor, November 29, 1918. Presented before THE INSTITUTE OF RADIO ENGINEERS, New York, December 11, 1918.

[1] Hazeltine, L. A., "Oscillating Audion Circuits," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 63, 1918.

[2] Langmuir, I., "The Pure Electron Discharge and Its Applications in Radio Telegraphy and Telephony," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 3, number 3, page 261, 1915.

[3] Miller, J. M., "A Dynamic Method for Determining the Characteristics of Three-Electrode Vacuum Tubes," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 141, 1918.

Vallauri,[5] and van der Bijl.[4] The object of the present paper is to display the general character of the variation of tube constants with the variation of grid or controlling potential, and also to describe additional methods for the measurement of three very important constants.

CONSTANTS OF VACUUM TUBES

It may be of interest at the outset to enumerate briefly and define the most important tube constants. This data is contained in tabular form in Table I, the symbols and dimensions being given together with the names of the persons to whom their definition is due.

TABLE I

Factor	Defined by	Symbol	Dimensions
Voltage Amplification	van der Bijl	μ	$\frac{dE_p}{dE_g}$
	Miller	k	$\frac{dE_p}{dE_g}$
Internal Impedance	van der Bijl	R_o	$\frac{dE_p}{dI_p}$
Mutual Conductance	Hazeltine	g	$\frac{dI_p}{dE_g}$
	van der Bijl	s	$\frac{dI_p}{dE_g}$
	(This Paper)	ρ	$\frac{dI_p}{dE_g}$
Detection (without grid condenser)	" "	D	$\frac{d^2I_p}{dE_g^2}$
Detection (grid condenser)	" "	σ	$\frac{d^2I_g}{dE_g^2} \cdot \frac{dI_p}{dE_g}$

[4] Van der Bijl, H. J., "The Theory and Operating Characteristics of the Thermionic Amplifier," PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, number 2, April, 1919.

[5] Vallauri, Professor G., "Sul Funzionamento dei tubi a vuoto a tre elletrodi audion, usati nella radiotelegrafia," "Estratto dal Giornale L'Elettrotecnica," January 25, number 3, and February 5, number 4, 1917.

COEFFICIENT OF VOLTAGE AMPLIFICATION.—This very important factor is an index of the relative effects on the plate current, of the grid and plate potentials. Specifically it is the ratio, $\frac{E_p}{E_g}$. In Langmuir's equation for the thermionic current in the plate circuit

$$I_p = A\,(E_p + k\,E_g)^{3/2} \tag{1}$$

it occupies a prominent place, the k being used as a coefficient to the term E_g, representing the grid or control potential. Dr. H. J. van der Bijl[5] has represented the plate current by another expression, as follows:

$$I_p = a\,(\gamma E_p + E_g + \varepsilon)^2 \tag{2}$$

In this case, γ is the coefficient of his extremely useful conception of the "stray field" and it is obvious that its reciprocal is of the same nature as the k in Langmuir's equation. The value of the voltage amplification constant depends upon the geometry of the tube, increasing as the ratio of the plate-filament, grid-filament distances and inversely as the spacing between grid wires. It is also dependent, as will be shown later, upon the region of the characteristic surface in which operation takes place. This latter relation would seem to make the utilization of the term *factor* preferable to that of *constant*. The manner in which the definition of this factor may be deduced from Vallauri's equation has already been indicated by Miller.

A method of measuring the amplification factor by means of direct currents is outlined in Dr. van der Bijl's paper; also a very useful method has been described by Dr. Miller.[3]

INTERNAL IMPEDANCE.—The value of the amplification constant being known, the hypothetical voltage operating in the plate circuit may be calculated as a function of the controlling potential on the grid. If there are no extraneous constants in the plate circuit, the plate current may be calculated by dividing this hypothetical emf. by a factor called the *internal impedance* of the tube. If in the plate circuit other constants are connected, the current may be calculated by the aid of a simplifying theorem suggested by Miller which reduces the plate circuit to a simple series circuit containing the plate resistance *internal impedance*, the added constants in the output circuit, and an operating emf. of value, $E_p = \mu\,E_g$. This simplification is of great value since it places the design possibilities of the

vacuum tube circuit on an equal basis with other branches of electrical and radio engineering. It is at all times desirable to design the circuit to fit the characteristics of the tube to be used.

The internal impedance is of the nature of a pure resistance at low frequencies. At radio frequencies, the plate-to-filament capacity of the electrodes, partially subordinates the effect of the internal impedance by acting across it as a branch circuit. This is of particular moment in the design of resistance-coupled amplifiers for short wave length ranges. Since the methods of measurement generally used in determining the magnitude of this factor are based upon audio frequency supply, it will not be necessary to take this into consideration at this time.

The measurement of the internal impedance has been covered by the methods of van der Bijl amd Miller. The first of these methods is based upon the fact that the internal impedance is the slope, $\frac{E_p}{I_p}$ of the d. c. characteristic. It may, therefore, be measured graphically from a plot of this relation, or by means of direct currents. Miller's method is a dynamic one, and is swifter and more accurate. The equation for the computation of R_o from the adjustments, however, involves the factor μ in a very important place, which means that the precision obtainable depends upon the measurement of μ. This is undesirable, not only for this reason, but also because a direct measurement of R_o only cannot be made without first determining the value of μ by a separate balance. These objections are not of a serious nature, but are nevertheless to be considered in the selection of methods for an extended research on vacuum tube problems.

The writer has employed a dynamic null method for the measurement of the internal impedance, which partially overcomes the above objections and gives much better minima in the indicating telephones. The circuital arrangement of apparatus is shown in Figure 1.

This is virtually a Wheatstone bridge connection. The audio frequency alternating current is introduced at the terminals of the slide wire, R_1 R_2. The resistance, R, may be made adjustable if desired and for maximum sensitiveness should be approximately equal to the internal impedance of the tube under measurement. The internal impedance of vacuum tubes may range from 5,000 to 500,000 ohms, and particularly at the higher values it may be advantageous to employ resistances

at R made by a spluttered film process. Specially wound Curtis coils may also be used and may be depended upon to be practically free from inductance. This matter becomes of importance only for the higher audio frequencies. The method of indicating the null point is of interest. In the writer's set-up

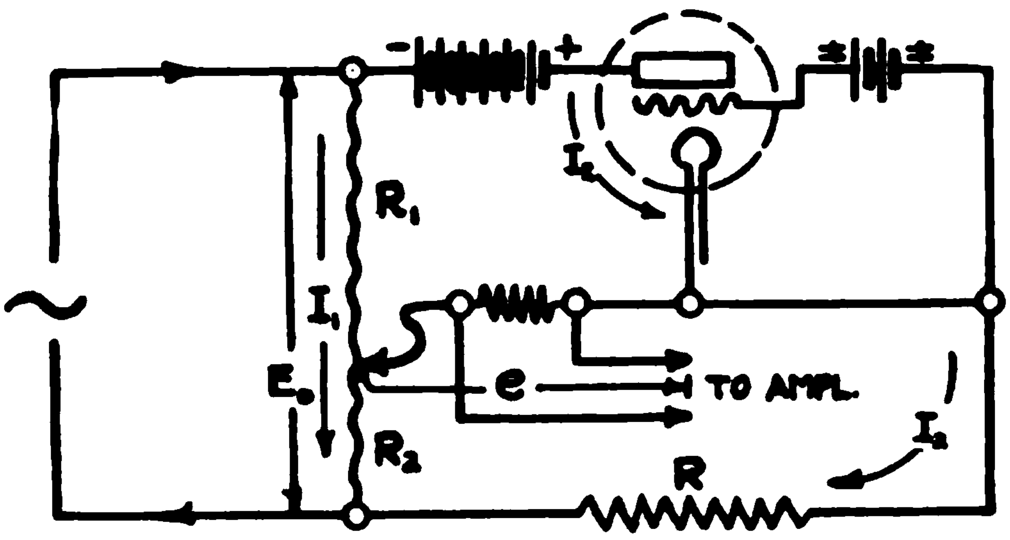

FIGURE 1

a small resistance of known value was inserted in the balance lead, the drop of potential being amplified by a two-stage vacuum tube system. It was found that this arrangement gave audibilities equal to those obtained with telephones inserted directly into the circuit, and possessed the additional advantage of reducing the extraneous inductance and resistance in the circuit to a minimum. The effect of the telephones in the measurements to be described was of the order of several per cent; for this reason the complication seems justified.

The theory of the arrangement shown is very simple. The alternating voltage, E_o operating across the terminals of the slide wire, causes two branch currents to flow, I_1 and I_2 as marked in the figure. For the condition of silence in the telephones, e and the current in the balance circuit containing the indicating apparatus must be zero. This means that:

$$R_2 I_1 = R I_2 \tag{3}$$

but

$$I_1 = \frac{E_o}{R_1 + R_2} \quad \text{and} \quad I_2 = \frac{E_o}{R_o + R} \tag{4}$$

Substituting into (3) gives:

$$\frac{R_2}{R_1 + R_2} = \frac{R}{R_o + R} \tag{5}$$

from which

$$R_o = \frac{R_1}{R_2} R \tag{4}$$

The computation of the internal impedance, R_o, from this relation is very simple and involves nothing but the ratio $\frac{R_1}{R_2}$ and the resistance, R. The main advantage, as mentioned above, of this method is the ability to obtain perfect null points. With ordinary care in the distribution of conductors it is readily possible to obtain balance points of a degree of ambiguity not exceeding 0.2 of one per cent. The measurement under these conditions becomes a matter of some precision.

MUTUAL CONDUCTANCE.—The ratio between the plate current and the corresponding grid or control potential is of great importance in determining the figure of merit of the device as an amplifier and oscillation generator. As pointed out by Hazeltine, it is of the dimensions of a conductance, and was therefore termed the *mutual conductance*, and represented by the symbol g. It seems to the present writer that the use of a symbol of this sort, having general engineering utility is undesirable and leads to confusion, and suggests the use of the Greek letter "rho" (ρ) for this purpose. The mutual conductance is evidently related to the other tube constants just defined by the expression:

$$\rho = \frac{\mu}{R_o} \tag{5}$$

and may, therefore, be computed from a knowledge of the factors involved. It is the slope of the "static" characteristic of the tube, plotted on the basis of I_p versus E_g (plate current versus control potential). It may be evaluated graphically by taking the slope of the tangent at any point on this curve. Since this method is laborious and may easily lead to considerable error, it may be better to measure the amplification constant and internal impedance separately and compute the value of ρ from the relation (5) above. While this is the better of the two methods available, it is still an indirect one, and it is regrettable that the measurement of this most important tube constant should be accompanied by so much labor.

The writer has developed a dynamic method for measuring ρ which retains the recognized advantages of null methods and possesses the additional one of giving the value of the mutual conductance directly. The method becomes

particularly simple if the slide wire, R, is calibrated directly in terms of the mutual conductance represented.

Dynamic Method for the Determination of the Mutual Conductance

The circuital arrangement of apparatus is shown in Figure 2.

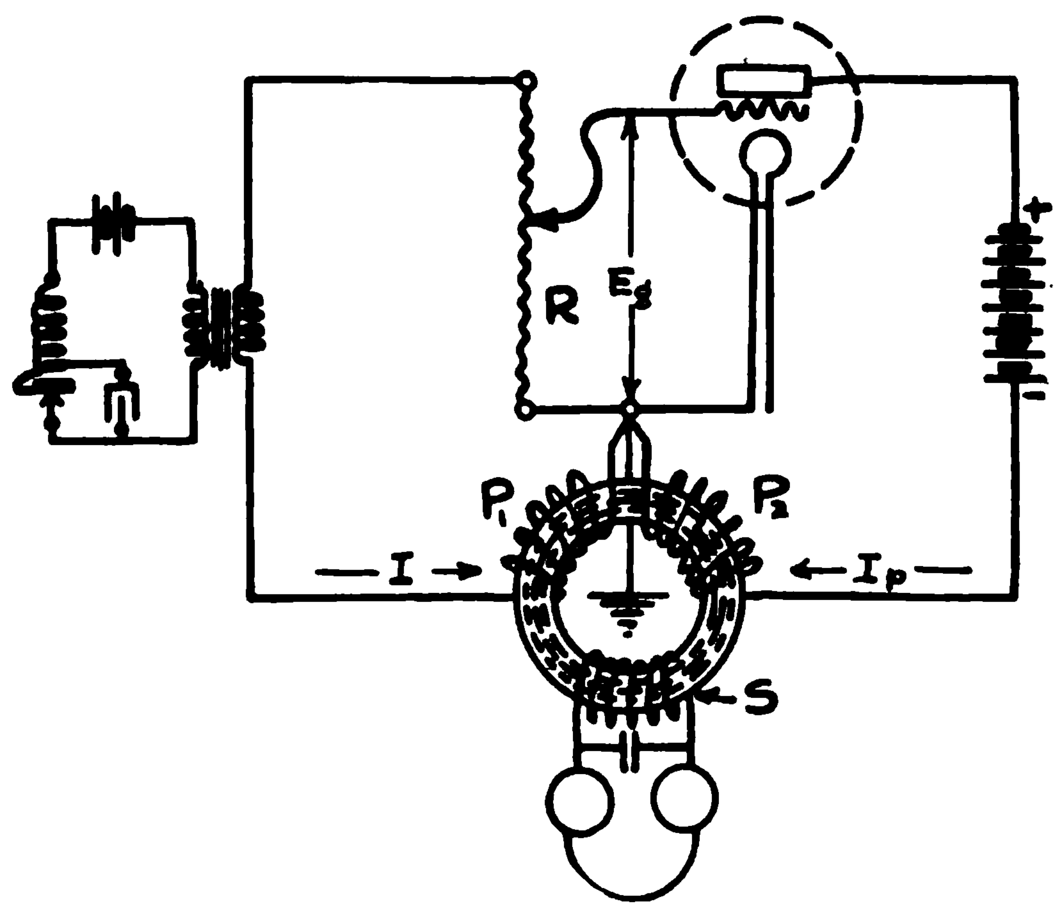

Figure 2

The alternating current for the measurement is supplied to the slide wire, R thru the primary, P_1, of the transformer. This may be of any frequency and may be supplied from an alternator, vacuum tube or buzzer source of the type shown in the figure. If the buzzer arrangement is employed, it is usually necessary to shunt the contacts with a large condenser in order to suppress undesirable harmonics tending to distort the secondary current from the true sine form. The leads to the apparatus from the generator should also be covered with metal sheathing and thoroly grounded, and also arranged so that the induction into the grid circuit will be at a minimum. The core of the transformer and the filament circuit of the tube are inter-connected and grounded in order to minimize residual sound at the balance point.

In the circuit shown, the input and output circuits of the tube are coupled by means of the toroidal core transformer, P_1 P_2 S, to a tertiary circuit containing the indicating apparatus. The theory of the arrangement is very simple.

If the leakage in the transformer is negligible, as will be the case if the core is a torus of the form shown in Figure 2, the induction into the tertiary circuit from the grid circuit primary will be:

$$E' = k\,t_1\,I \tag{6}$$

where k is a constant of little interest. Similarly from the plate circuit we have the induction:

$$E'' = k\,t_2\,I_p \tag{7}$$

If now, the windings are so connected that the two emfs. are vectorially opposed, the resulting emf. operating in the circuit will be:

$$E' - E'' = k\,(I\,t_1 - I_p\,t_2) \tag{8}$$

The indication being taken as the condition of silence in the telephones (or amplifier) connected in this circuit, this state of affairs leads to:

$$I\,t_1 = I_p\,t_2 \tag{9}$$

The value of the plate current depends upon the grid voltage as indicated by the relation, $I_p = \rho E_g$, and

$$I_p = \rho E_g = \rho\,R\,I \tag{10}$$

so that

$$t_1 = \rho\,R\,t_2 \tag{11}$$

and finally

$$\rho = \frac{t_1}{t_2\,R} \tag{12}$$

which defines the value of the mutual conductance, ρ, in terms of the resistance, R, across the grid circuit. For practical reasons, it is well to keep the value of R required for normal measurements, at a low value of the order of 100-1,000 ohms. This matter may be adjusted nicely by selecting a suitable ratio t_1/t_2, such as, for instance, 10:500. If the ratio is kept at a constant value, the relation becomes very simple, permitting the calibration of the slide wire, R, directly in terms of ρ. The method then becomes both swift and accurate and should be particularly useful in connection with acceptance tests of vacuum tubes intended for use as amplifiers and oscillators where a high value of ρ is desirable. The latitude of tolerance may be marked directly upon the slide wire thus reducing the mental labor of such tests considerably. The whole matter then becomes a mechanical proposition, and may be safely intrusted to inex-

perienced operators. Measurements on certain tubes by this method will be described in a later paragraph.

DETECTOR CONSTANTS.—In addition to its use as an amplifier and oscillator, the vacuum tube has a distinct field of utility as a detector of radio frequency oscillations. When used for this purpose, two methods of operation are possible: (a) without grid condenser but with biasing battery in the grid-filament circuit, and (b) with grid condenser. The phenomena attending the latter mode of functioning have been well described in the classic paper by Armstrong[6] as well as in numerous other publications.

In general, when the grid condenser is omitted, operation takes place upon a region on the characteristic surface

$$I_p = \phi\,(E_g, E_p)$$

which is curved or bent, the indication in the telephones being the result of a variation in the mean value of the plate current due to the superposition of the radio frequency oscillations. Two regions of operation are possible, as shown by the static characteristic curves of Figure 5. It will be observed that the second derivative of I_p with respect to E_g, which determines the curvature, becomes appreciable at two points. One of these represents saturation and involves considerable space current, the other occurs when the curve is leaving the axis, $I_p = 0$. The latter region is usually the best operating point on account of greater curvature.

The curvature of the static characteristic curve at either of these points is a direct index of the detecting merit of the tube. The truth of this proposition may be readily established as follows:

Assume that the curve shown in Figure 3 represents the plate current as a function of the grid or controlling potential, this being the graph of some function:

$$I_p = \phi\,(E_g) \tag{13}$$

The operation is to be performed around some point on this curve, such as E_g marked in the figure. An increment equal to $\Delta\, e_g$ is given the control potential, followed by an equal one of opposite sign. We have then:

$$\begin{aligned} I_p + \Delta_1 I_p &= \phi\,(E_g + \Delta\, e_g) \\ I_p - \Delta_2 I_p &= \phi\,(E_g - \Delta\, e_g) \end{aligned} \tag{14}$$

[6] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 3, number 3, September, 1915.

The response in the telephones is proportional to the variation of the mean value of the change in plate current, or

$$\frac{\Delta_1 I_p - \Delta_2 I_p}{2} \tag{15}$$

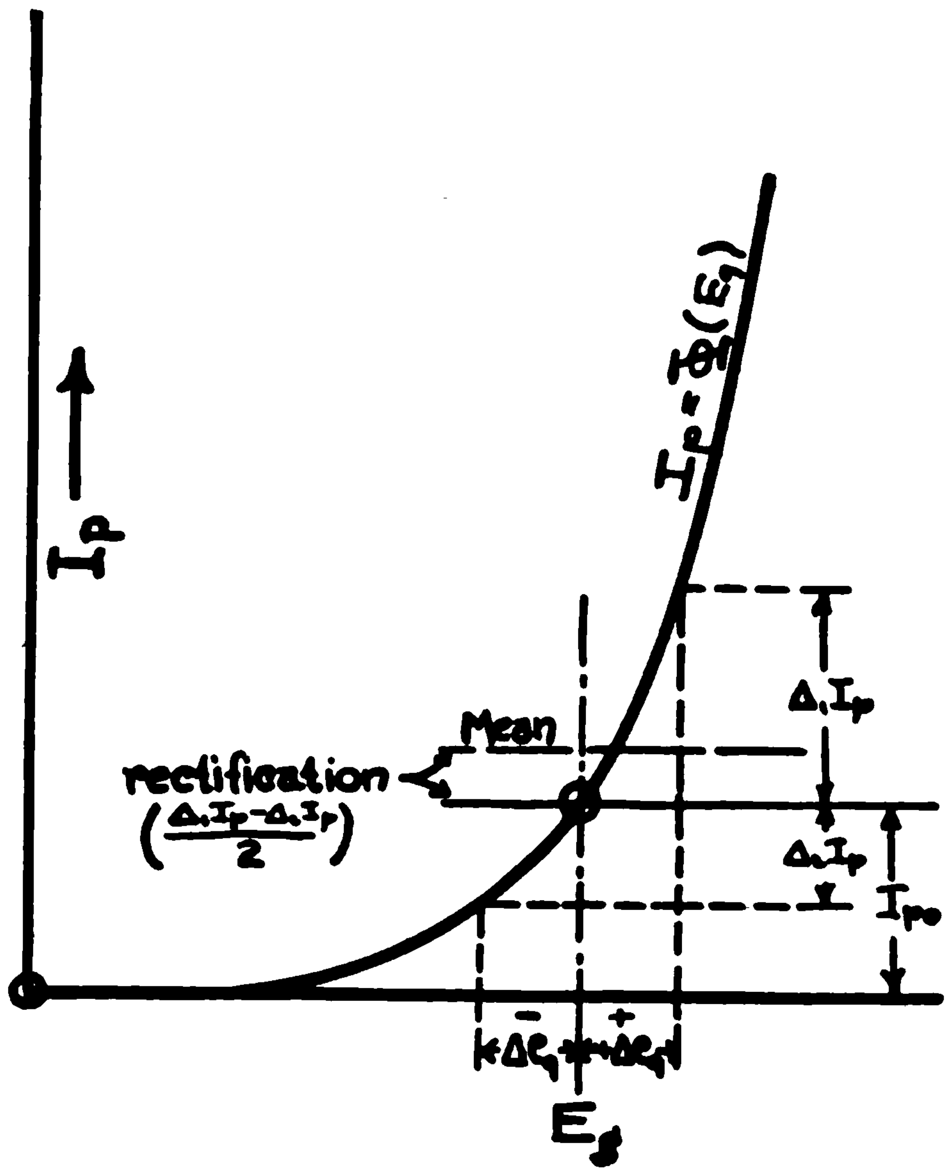

FIGURE 3

from the steady state value, I_p. For a given stimulus, Δe_g, the rectification index is:

$$\frac{\Delta_1 I_p - \Delta_2 I_p}{2 \Delta E_g} \tag{16}$$

or is proportional to the difference in the slopes of the curve at the points corresponding to the constraints of operation. The function being continuous, assuming the increments to be infinitesimal and going to the limit this ratio becomes:

$$D = \frac{d^2 I_p}{d E_g^2} \tag{17}$$

or for small variations of E_g the rectification depends upon the second derivative of the plate current with respect to the grid potential. This, it will be remembered, is also the derivative of the mutual conductance with respect to E_g. The value of

this differential is fundamental in determining the merit of the device as a detector without grid condenser, and may be called the *detector constant without grid condenser*, and denoted by the symbol, D. It may be readily determined graphically from the curve of mutual conductance as a function of the controlling potential, or computed from measurements of the internal impedance and amplification constant at two points as close together as possible. These methods are open to the objection of being indirect, but a suitable direct method does not seem to be available.

(b) OPERATION WITH GRID CONDENSER.—As is well known, operation of the tube with a series condenser in the grid circuit involves the rectification of the impressed oscillating emf. on the grid to produce a charging current thru the condenser which is uni-directional and hence charges the grid condenser continuously to produce an increasing negative potential on the grid. The radio frequent voltage is super-imposed upon the changing mean potential of the grid, and the point of operation instead of remaining fixed upon the static characteristic curve, slides down along the curve and produces a variation of the mean plate current upon which is superposed the radio frequent changes. On this basic theory we are obviously interested in two changes; first, the rectification in the grid circuit, and second, upon the slope of the $\frac{I_p}{E_g}$ curve. A definition of the merit of the tube used in this manner must contain both of these factors. The first effect, in the light of the previous discussion, we may expect to be dependent upon the second derivative of the grid current-grid potential curve. The second is obviously determined by the value of ρ. Superimposing these effects, an appropriate definition of the detecting action would seem to be:

$$\sigma = \frac{dI_p}{dE_g} \cdot \frac{d^2 I_g}{dE_g^2} \tag{18}$$

In some tubes, the region of maximum curvature in the grid current-voltage curve may correspond to working points involving finite positive values of grid potential, in which case it is desirable to use a "grid-leak" resistance in order to place the starting point at the proper place on the characteristic curve. This does not effect the definition, however, and is of very little interest here. It is also to be noted that as the oscillation persists and the charging of the grid condenser continues, the

region of operation shifts, producing simultaneously a change in the curvature of the $\frac{I_g}{E_g}$ curve so that the above definition cannot be regarded as complete except when the stimulus is weak and of short duration or a grid leak is used and the equation of the grid current as a function of the impressed grid emf. may be placed in constant exponential form. Otherwise, the rectification is no longer a function of the curvature of the characteristic alone, but also depends upon the magnitude of the initial oscillatory pulse. In short, the precise nature of the phenomena attending the operation with the grid condenser being complex and imperfectly understood it may be well to bear in mind the possibility of discounting the value of the above definition. Its main value lies, however, in its usefulness in indicating the general possibilities of the tube as a detector used in this connection.

The value of the detecting constant with grid condenser, denoted by the symbol (σ) may be determined by the aid of a dynamic method similar to that used for the measurement of the internal impedance. The circuital arrangement is shown in Figure 4.

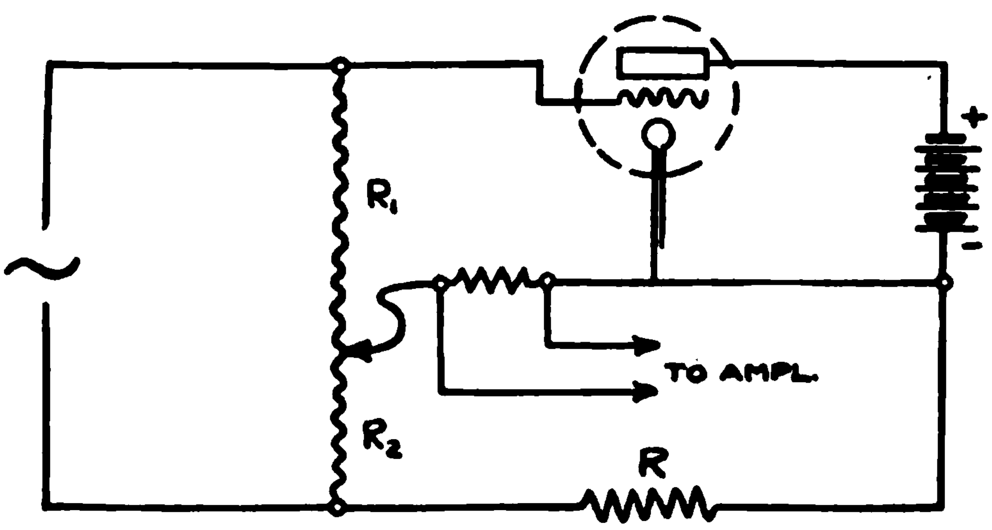

FIGURE 4

For the condition of silence, as in the previous case, we have for the effective resistance of the grid circuit of the tube:

$$R_g = \frac{R_1}{R_2} R. \tag{19}$$

The reciprocal of this quantity, represents a conductance and determines the slope of the grid current-voltage curve. If two measurements are made at points not widely separated, of the reciprocal (g), substitution in the expression:

$$\frac{g' - g''}{e_g' - e_g''} = \frac{R_g'' - R_g'}{R_g' R_g'' (e_g' - e_g'')} \tag{20}$$

yields an approximate value of the second derivative, which when multiplied by the value of ρ corresponding to the mean value of the grid potential taken in the preceding measurements, will give the approximate experimental value of the detecting constant. The value obtained in this manner, while not theoretically accurate, is nevertheless of more value from a practical viewpoint since operation never takes place over a region which is theoretically infinitesimal. The method is perfectly simple and straightforward, but for the reasons cited above, the value of the result as a final index of the operation is questionable.

Experimental Investigation of the General Relation Between Tube Parameters

The summarization of definitions and methods of measurement contained in the preceding paragraphs may be advantageously applied to an investigation of the general character of the relations between the constants of vacuum tubes. In the following paragraphs will be reproduced some curves which have been experimentally obtained with the methods described above and which have been made upon a typical, widely used form of thermionic amplifier. Dr. Miller in illustrating the use of his methods, has already published some curves connecting the amplification coefficient and internal impedance with the plate voltage. The use of this independent variable gives a result, however, which is only a very small part of the whole story. Information of much greater value may be obtained by plotting the variation of the tube parameters in parallel with the static characteristic, or against E_g taken as the independent variable with E_p as parameter. A complete investigation for a given tube would involve the repetition of this process for various values of filament temperature. This introduces a total of three independent variables and involves considerable experimental labor, but this would be compensated for by the completeness of the study and fertility in deductions made possible. A complete and logical research might be planned upon this course of procedure with each tube, by varying the geometry of the electrode system, that is to say, by varying the grid-filament spacing with plate-filament spacing being regarded as a parameter. It is unnecessary to observe that a complete investigation undertaken along these lines, while involving considerable time and experimental labor, would permit of a correlation of structural methods and results which would be of the greatest value in the future design of vacuum tubes.

The general character of the variation of the tube constants defined above with the grid or controlling potential (E_p being taken as a parameter), is of primary interest in the study of a certain type of tube. In order to obviate the necessity for repeating this work with different values of filament temperature, one temperature should be selected for the work which is *optimum,* and represents the maximum value of the product of life and desirability, the latter being defined by the increase in the important constants. In the curves to be displayed, an investigation having previously been made on this point, a filament current was used which was considered to be an optimum value.

The family of *static* characteristic curves in Figure 5 are representative of the tube used.

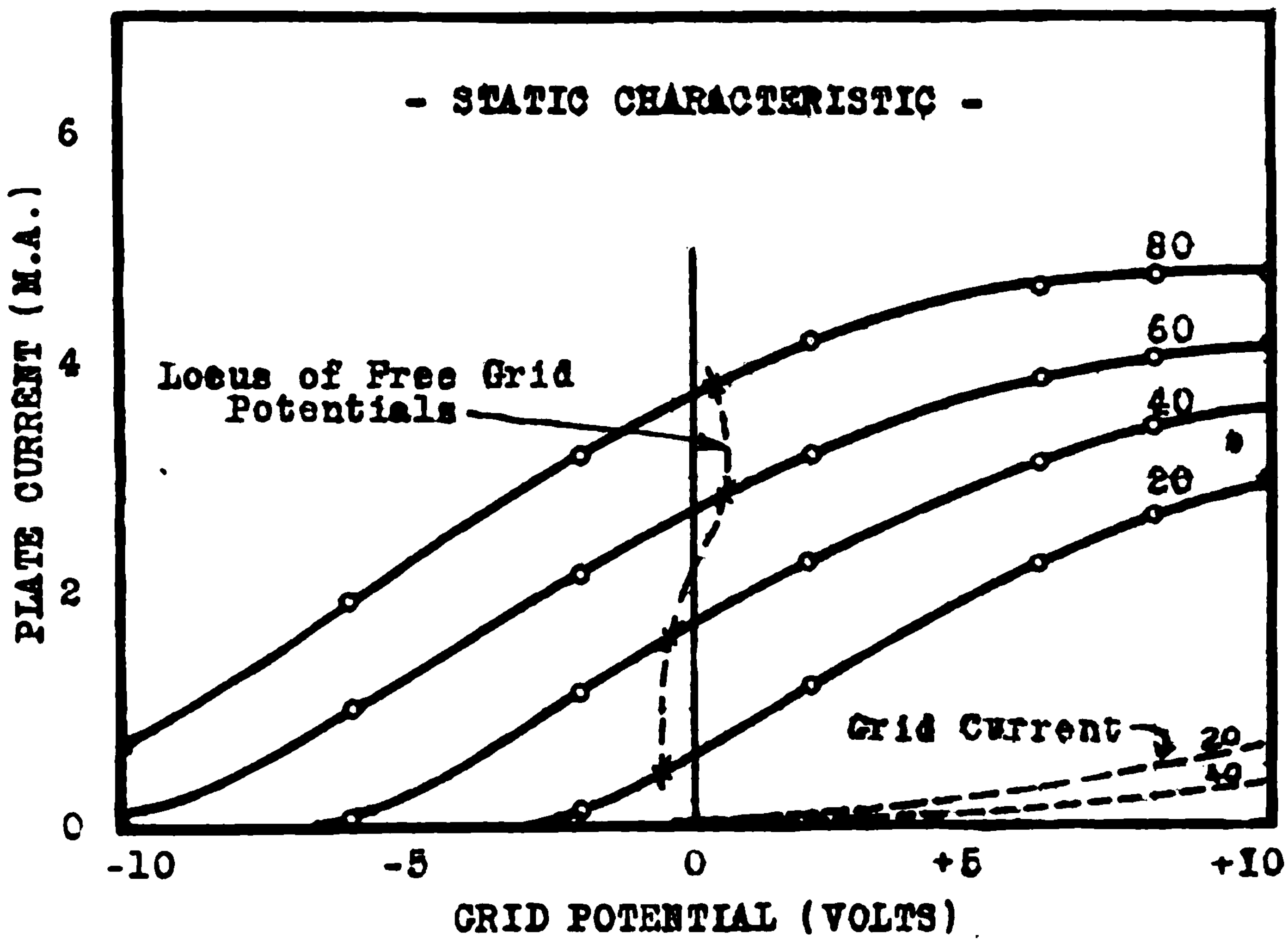

FIGURE 5

In this figure, two of the grid currents have been represented by the dotted curves. The free grid potentials or the potential which must be applied to the grid in order that the plate current may have a value identical with that which is obtained when the grid is disconnected, are represented by the dashed curve marked "locus of free grid potentials." This is a matter of secondary interest, but is sometimes helpful in making deductions concerning the operation of the tube.

The amplification coefficients were measured with the same

values of plate potential and with the same independent variable, E_g. Miller's method proved very useful for this purpose. Great care is necessary in the arrangement of apparatus to reduce electrostatic induction effects which tend to obscure the true balance point. The value of such initial care in the set-up of the apparatus is well shown by the agreement between the final smoothed curve and the observation points. The results are depicted in Figure 6.

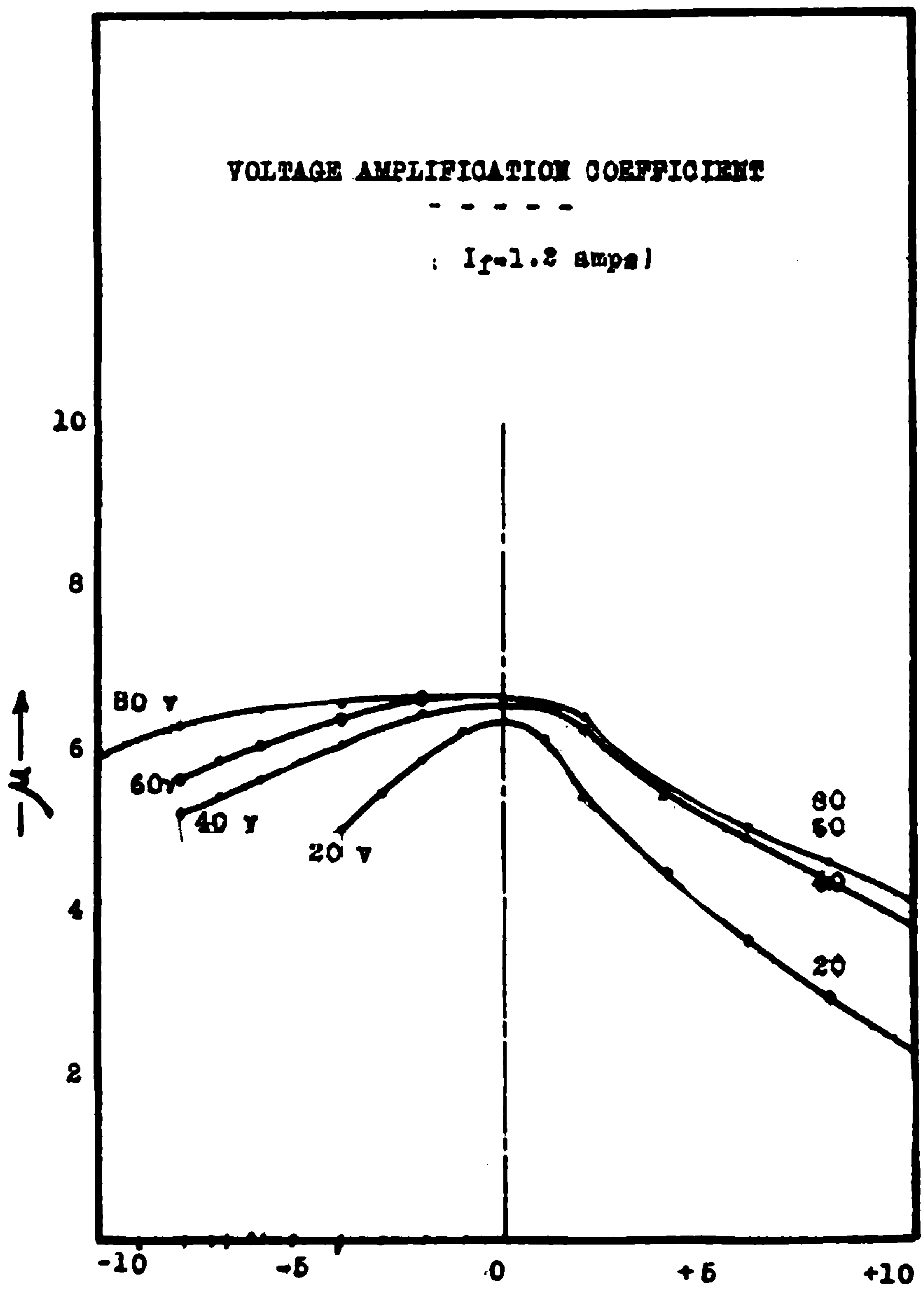

FIGURE 6

The salient feature of these curves lies in the fact that the amplification factor is not constant, as has been intimated from measurements with direct currents by Dr. van der Bijl, but

depends greatly upon the grid potential in the region in which operation takes place. The falling off for positive values of grid potential and smaller plate voltages is particularly noticeable. The reason for this is not apparent, *a priori*, from the theory, since the relation between the effects of stray field and the grid potential would seem to be constant and independent of any value that the grid potential might assume. However, the equation upon which the theory rests represents conditions only within limited constraints on the characteristic surface, particularly in the region where the plate current is small and the grid potential is negative or near zero. After the point of inflection on the curve is passed, the expressions of Langmuir and van der Bijl are not even correct in form, so that a wide discrepancy between theory and the results is not seriously disturbing. The following explanation of the effect in question has suggested itself to the writer.

If the exact exponential variation of the plate current with respect to the inter-electrodic forces is assumed to be arbitrary, and the concept of the relative effects of the plate and grid potentials only is retained, we may represent the plate current as some function of these forces as follows:

$$I_p = f(E_p + \mu E_g) \tag{21}$$

The only matter of interest in this connection is the relative effects upon the plate current, of the grid and plate forces, since this determines the amplification of voltage in the tube. Assume that the distribution of forces and scalar potentials may be approximately represented by the diagram of Figure 7.

In this figure the inter-electrodic distances are represented by the abscissas, the ordinates representing the value of the scalar potentials, V. F is the emitting surface, and the dashed line represents the grid electrode. The positive potential is applied to the plate (P in the figure) and when the tube is cold the scalar potential may be represented by the straight line AB. In this case the grid is not connected. When the grid is connected (the practical case) the only field existing at G is that due to the plate potential acting thru the grid wires as pointed out by Dr. van der Bijl. This he has very appropriately termed the *stray field*. It is represented in the figure by the dotted line Ab. When the tube is hot and the surface F is emitting electrons, AB and Ab become roughly, as shown in the figure, the curves marked AB' and Ab', respectively. The curvature is due to the effect of the electronic atmosphere, which consti-

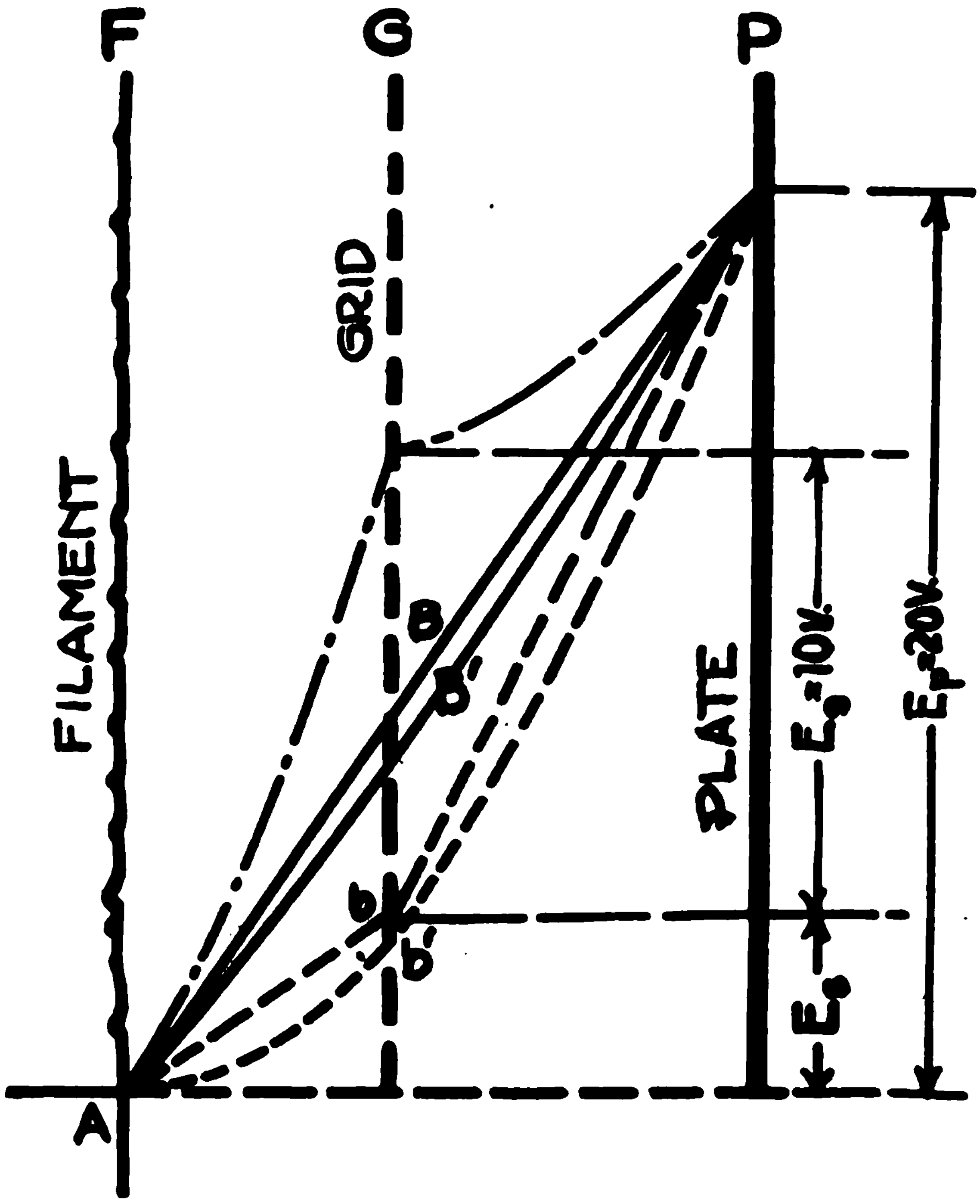

FIGURE 7

tutes a negative cloud. At the operating filament temperature the slope of the curve at the filament is finite and the saturation effect of the *space charge* is not reached when the grid is disconnected. When the grid is connected, however, space charge saturation between grid and filament is undoubtedly attained and the slope of this curve at the emitting surface is zero. The emission velocities at the surface of the filament are ignored as being inappreciable at a temperature of 1,300 degrees Kelvin, the point of operation. The reason for the slope of the potential function attaining a zero value when the emission is copious has been well brought out by Richardson,[7] with the aid of Poisson's equation. This is the initial state of affairs when the grid potential is zero. When a positive potential is placed upon the grid, the potential curves probably assume the general form shown in the figure by the dot and dash curves. In this case the vector potential, $\nabla \cdot V$, the force acting upon the

[7] Richardson, O. W., "The Emission of Electricity from Hot Bodies," Longmans, Green and Co., London, 1916.

electron is greater in the region between the grid and filament than in the plate-grid space, this increasing as the plate potential is decreased. There being now, more electrons in the plate grid space than before, the effect of space charge is beginning to be felt and the curve in this region sags slightly as shown. The force tending to overcome this condition is that due to the plate and for small plate voltages it seems readily possible for the slope of the potential curve at the grid to be zero. This means, of course, that there is no force operating upon the electron. The plate contributes nothing. This effect disappears at higher values of plate potential and the curve straightens out.

From the above statement of the condition of affairs in the inter-electrodic space, it would appear that when the potential on the grid is in the neighborhood of zero, and the grid-filament is enjoying the effects of space charge, the field intensity between plate and filament being high enough to attract strongly any electrons that manage to escape the congestion, we may expect the grid potential to have a greater effect upon the electronic flow than that of the plate. This carries with it the idea of a large value of μ from the relation (21) above. Also when the grid potential becomes positive with respect to the filament, the space charge effect mentioned is broken up and the conduction is increased. This is accompanied by a shift of the space charge from the filament-grid region to the plate-grid region, and, as noted above, the curve between the plate and grid electrodes sags to correspond. In this case there are plenty of electrons available at the grid plane, but the reactive force is the space charge. A slight increment in the plate force results in a partial neutralization of this effect by taking some of the electrons out of the space, so that now the plate potential becomes the predominating force since any increase in grid potential will obviously only increase the tendency to space charge. This results, of course, in a lower value of μ, which effect increases as the ratio $\frac{E_g}{E_p}$ decreases. The gist of the whole matter is that the space charge, with its accompanying saturation, shifts from one region to the other and changes to correspond, the relative effects of the grid and plate potentials and thus the value of μ.

The variation in the internal impedance is well shown by the curves of Figure 8.

The measurements represented were made with the aid of the

bridge arrangement described in this paper, the minima obtained with Miller's method being ill defined. The results are homogenous and very satisfactory. The salient feature of the curves is that the resistance of the tube undergoes extreme variation,

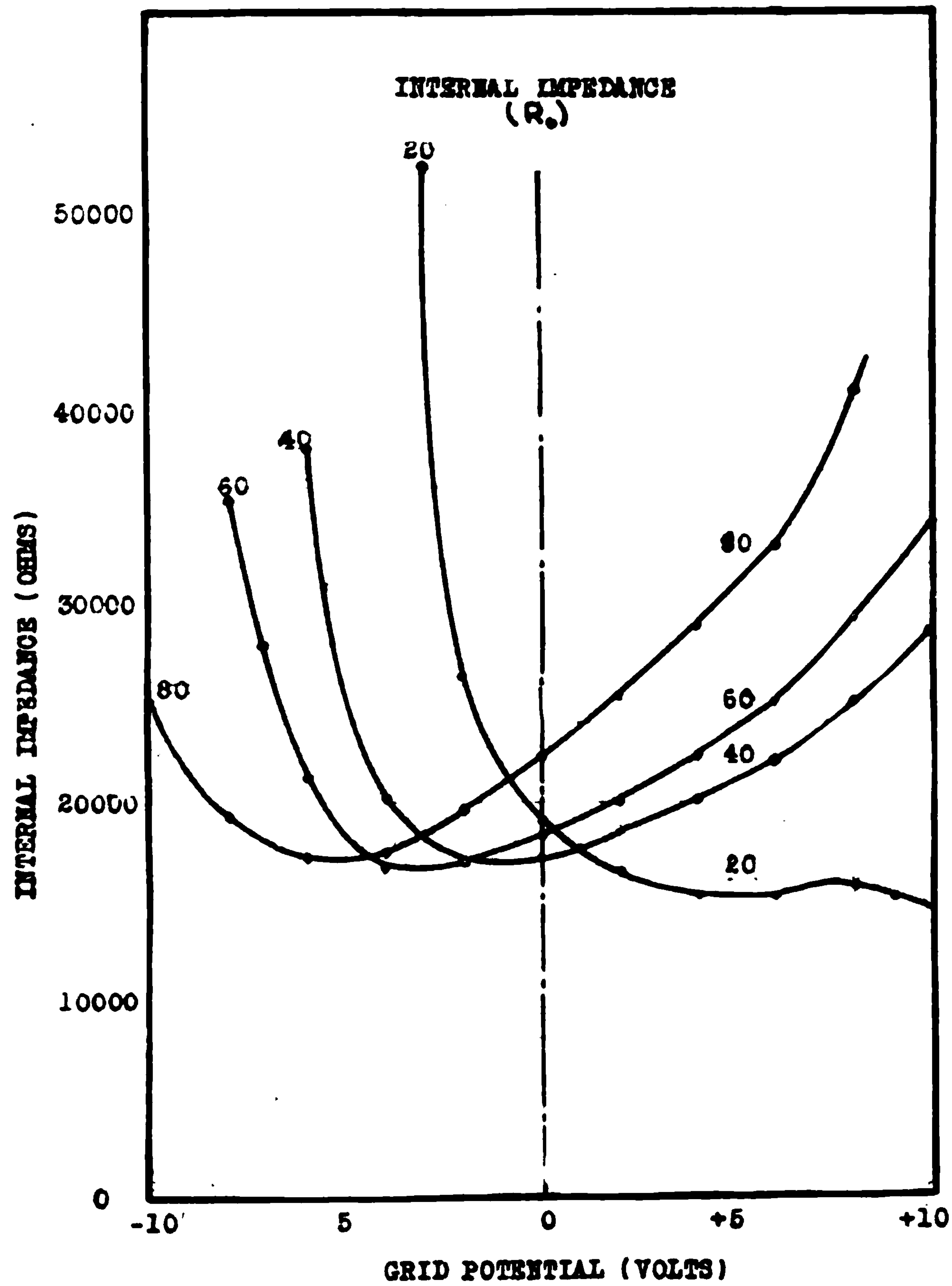

FIGURE 8

the minima representing roughly the points of inflection on the static characteristic curves.

The factor of prime importance in connection with the use of the tube as an amplifier or oscillator is the mutual conductance. This may be computed from the values of μ and R_o measured above from the relation (5) or it may be directly determined by the writer's method previously described. In Figure 9, the solid

curves represent the results of computation from the amplification constant and the internal resistance, while the observed points, using the method in question, are represented by the small circles. The agreement between the two methods seems to be satisfactory and the results are very interesting.

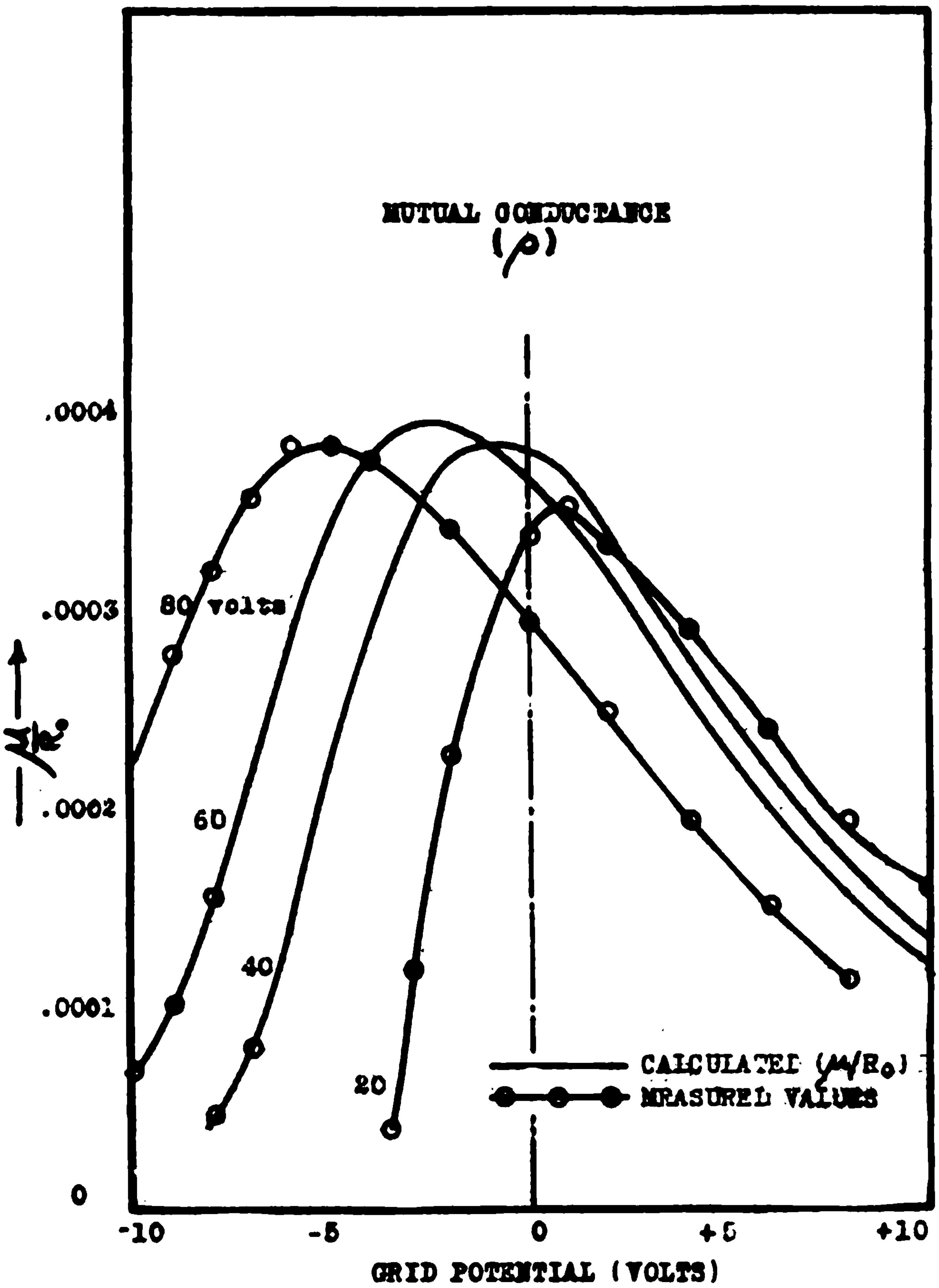

FIGURE 9

It will be noted that none of the maxima are very broad which shows that amplification at these points will be accompanied by distortion if an extended operating region is involved. Also when using the tube for amplification purposes in connection with radio receiving, the point of operation should be carefully

selected by means of a biasing potential of suitable value applied to the grid. In selecting the operating region for this purpose, other considerations are pertinent. For instance, the maxima are not equally desirable since their magnitudes are different, hence the optimum value of plate potential should also be selected. This should also be preferably situated on the left hand side of the zero center line, so that the grid or input circuit offers no conductance when connected across the terminals of the condenser or the output circuit of the preceding tube. Having determined by intelligent methods upon the proper point for the operation of the tube, we may now proceed to design the connecting link between the tubes which will introduce the optimum impedance into the output circuit and give the maximum voltage output across the input circuit of the next tube. In this very practical matter, we are aided considerably by a knowledge of the internal resistance of the tube. The design of amplifiers is a matter of some engineering importance and will be considered at length in another paper.

The detecting constant, D, for operation without grid condenser may be evaluated by taking the slope of the tangent to the ρ curve at any point. This is not an entirely accurate procedure but is the best method at present available. The results of this method applied to the tube used in this work are represented in Figure 10.

The truth of the belief that the better operating point corresponds to low values of plate current is well brought out by these curves. Also it will be noted that low plate voltages are favorable to such operation. From the point of view of plate battery economy this is an important point. In general tubes operate better as detectors with grid condensers, and in special cases, with grid-leak resistances in addition.

The statement just made concerning the operation of the tube with grid condenser emphasizes the importance of studying detector constants under these conditions. This is somewhat of a tedious process but the results are interesting. The Wheatstone bridge arrangement described may be used to measure the effective resistance of the grid filament circuit at various grid potentials. The reciprocal of this quantity is the grid conductance, $\frac{I_p}{E_g}$. In Figure 11 this conductance has been plotted against positive values of grid potential.

There is no conduction for negative values of E_g if the emission is pure (that is, there is no positive emission) and the tube

has been well evacuated, so that negative values of E_g have no interest. The slope of the conductance curve is evidently the second derivative of the grid current function with respect to the applied potential. This is one of the quantities entering

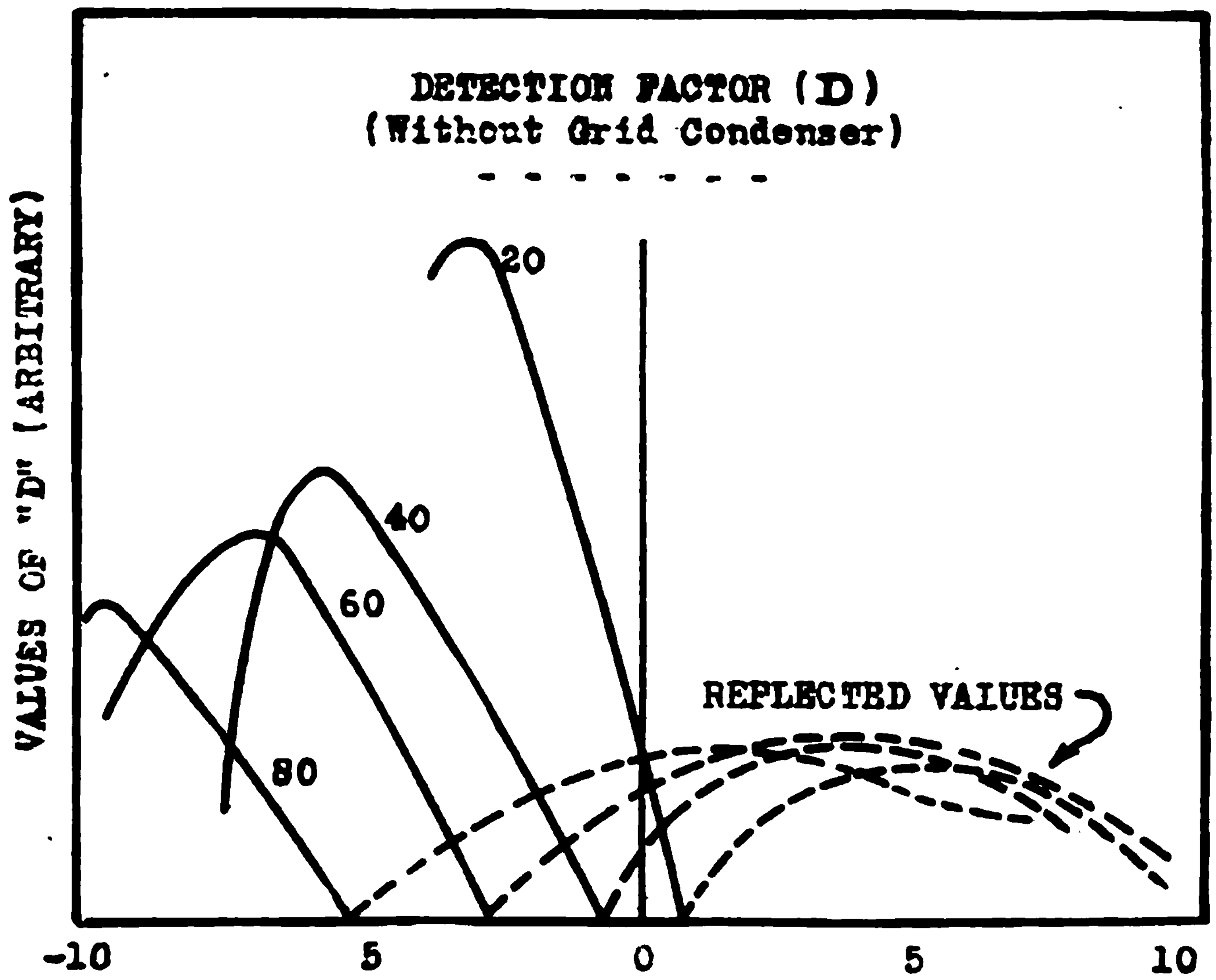

FIGURE 10

into the definition of the detecting constant with grid condenser. The dashed curves in Figure 11, represent the magnitude of the second derivative and have been derived from the conductance curves by taking the tangent at various points. It is interesting to note that the larger values correspond to the lower plate potentials. The complete definition of the detecting factor involves in addition to the grid rectification index, the slope of the $\frac{I_p}{E_g}$ curve, or ρ. The curves in Figure 12 represent the addition of this factor and exhibit completely the detecting action of the tube.

Several practical deductions may be immediately made from these results. In the first place, when the tube is to be employed as a detector (with grid condenser), a grid leak is desirable, especially when strong signals are to be impressed upon the system. This leak resistance should be placed between the

grid and the positive leg of the filament and should have a value necessary to place the starting point of the grid potential at a point on or above the maxima on the curves in Figure 12. A

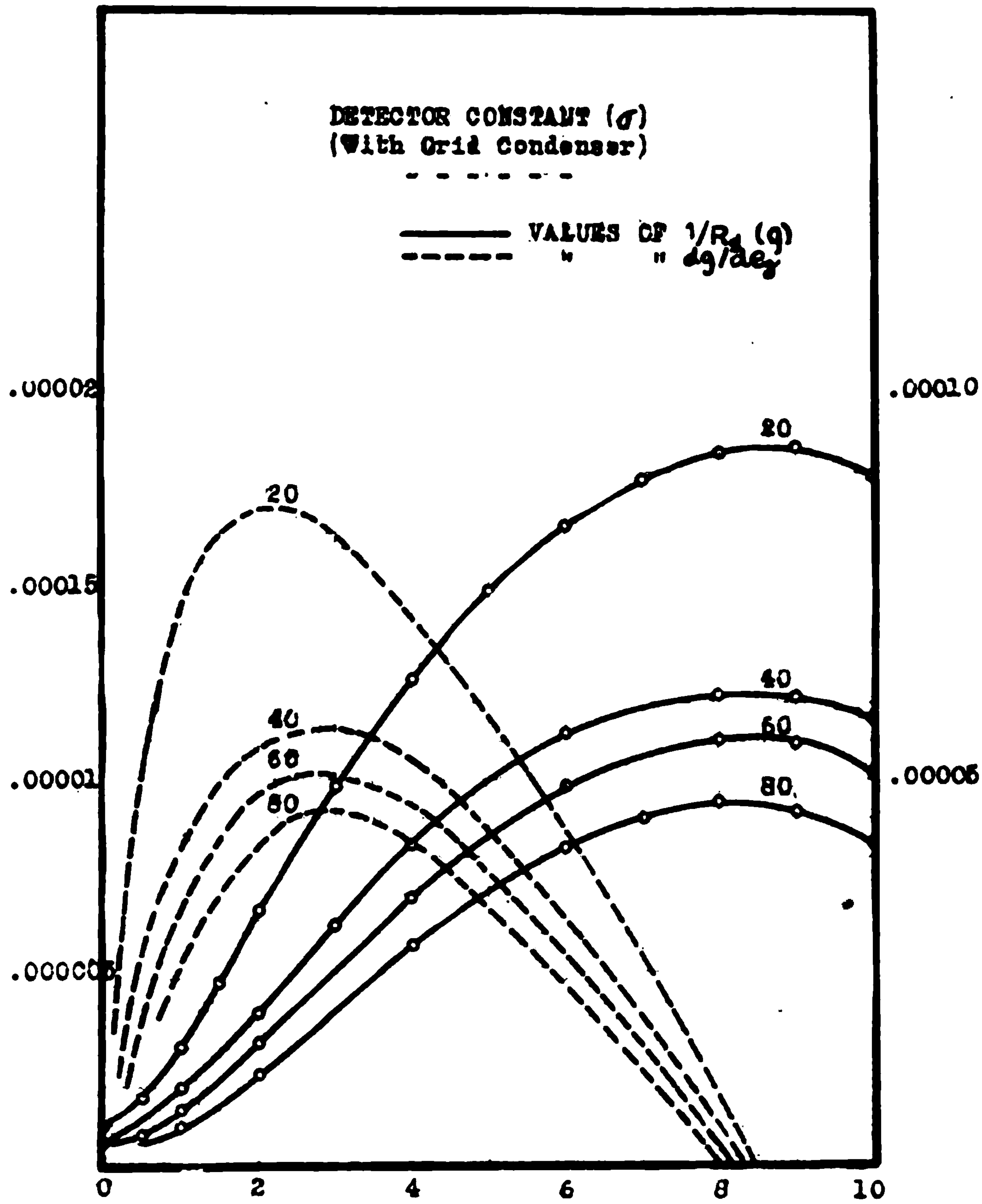

FIGURE 11

plate voltage of 20 or possibly less (this point not having been investigated) should be used, since this gives the greatest value of σ. In the case of weak signals, whether or not a grid leak should be employed is problematic, since theoretically when the grid current-voltage curve intersects the axis, $I_g = 0$ the curvature is infinite, and a singular point condition is obtained. The detecting factor is then infinite also, and the rectification is perfect. This will only be the case for very weak signals and for general purposes, it may be well to sacrifice if necessary, the response

for weak signals to gain efficiency on the stronger signals by employing a grid leak resistance. In the region between the zero grid voltage and the voltage corresponding to maximum values of σ, the change in σ is great. Similarly after the maxima have

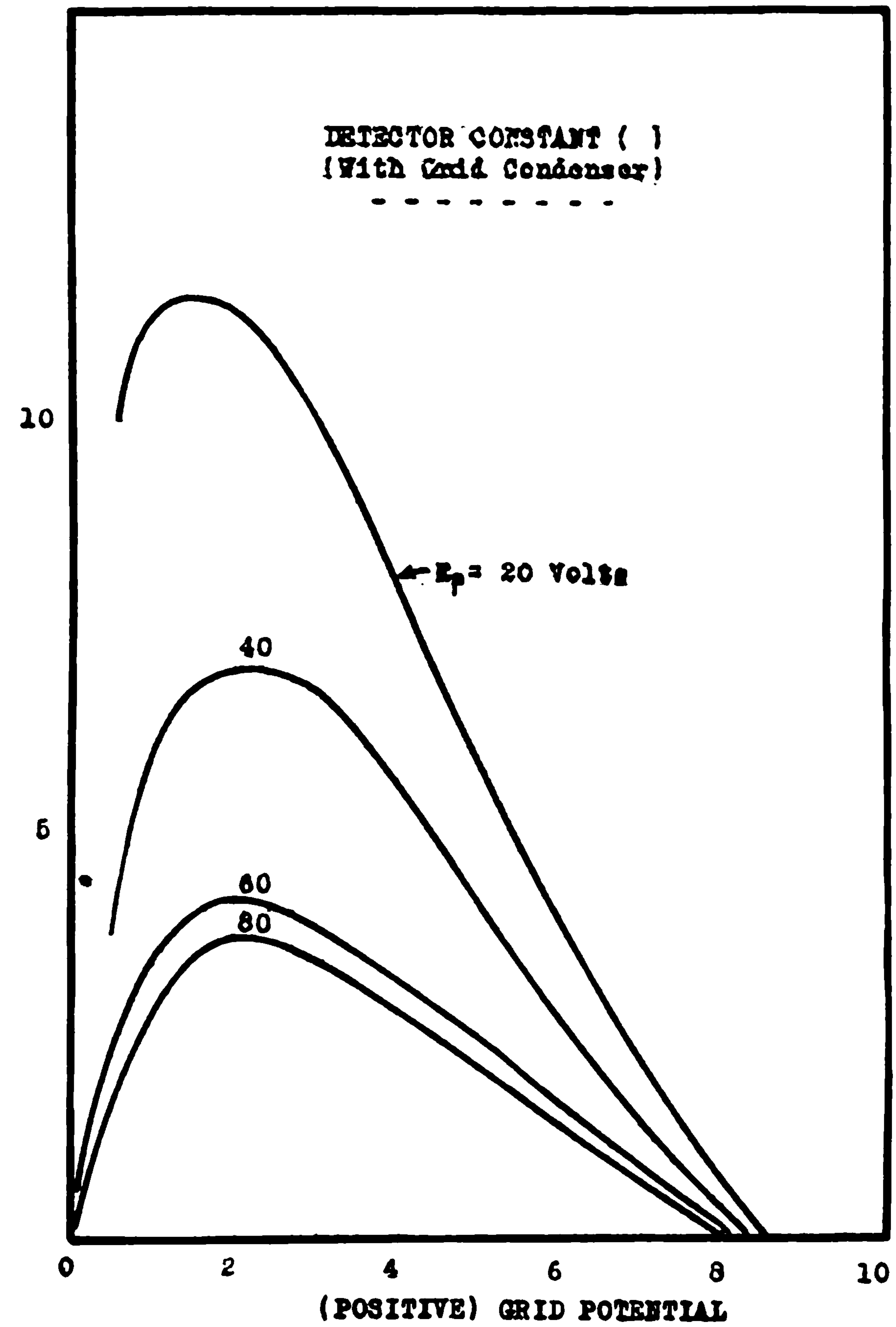

FIGURE 12

been passed, but in this case, the irregularity is not serious since the grid potential will shift to the negative when the charging starts, passing thru the maximum value and being augmented in the process. As a matter of fact, this consideration leads to the question as to whether or not it may be advantageous, in order to cover all possibilities in the matter of signal strength, deliberately to place the initial working point upon positive

values of E_g beyond the maxima. Just how far this may be carried out without sacrificing to too great an extent, the response on small stimuli, will be largely a matter of experiment.

Philadelphia, Pennsylvania,
November 24, 1918.

SUMMARY: After giving the definitions of various tube constants for amplification and detection and the methods of measuring them, the author discusses certain special questions.

Among these are grid condensers and grid condenser leaks, and the variation of the various tube parameters with tube construction and operating conditions.

ADDENDUM I*

In the paper a method for measuring the mutual conductance was described which involved the use of a toroidal core balancing transformer. Since this was written an alternative arrangement has been devised which may be used for this purpose with equal success, and which does not require the special transformer of the original arrangement. This method should therefore be more popular with those who do not care to undertake the somewhat tedious task of constructing a transformer with a winding of five hundred turns on an iron torus.

The circuital arrangement is shown in Figure A:

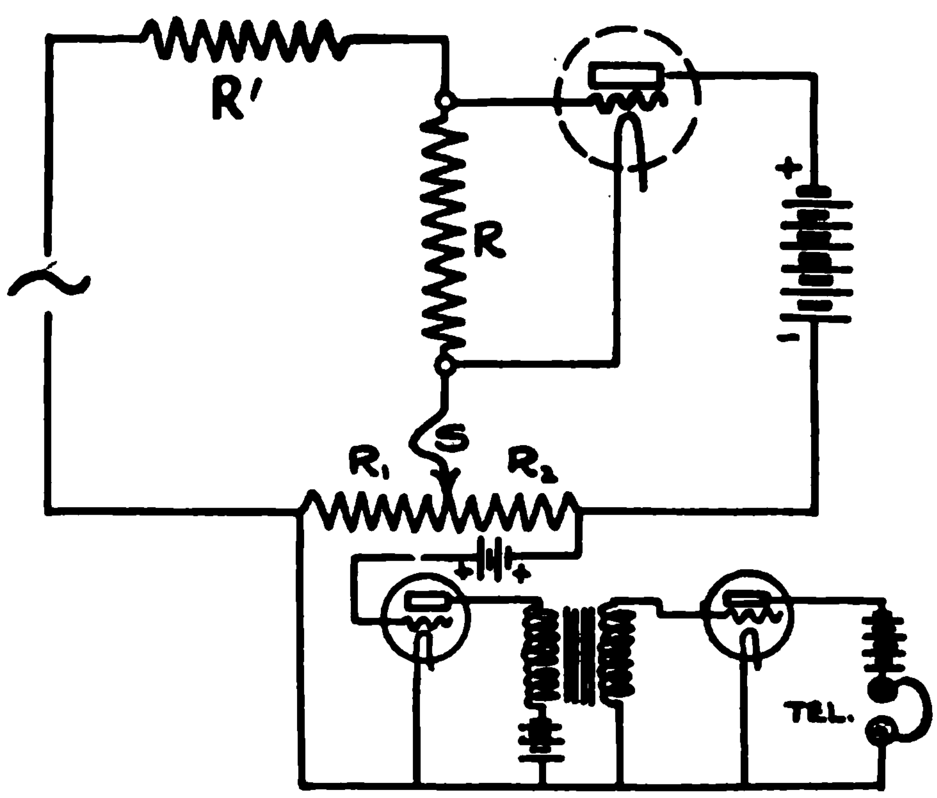

FIGURE A

* Received by the Editor, January 15, 1919.

Here, as will be noted, the transformer primaries, P_1P_2 have been replaced by the resistances, R_1 and R_2. The large resistance R' has been inserted into the supply circuit in order that the main current may not be greatly affected by the variation of R_1 in series. The resistance R_2 is variable and a part of the plate circuit, however, but being of the order of from 70 to 100 ohms with a possible variation of 30 ohms, its effect on the mutual conductance is quite ignorable. The resistance, R, across which the grid is connected is fixed at about 1,000 ohms, this value being chosen for the simplification of numerical computations of ρ from the settings.

The position of the sliding contact, S, is found for which the emf. across the terminals of the slide wire is zero and there is no indication in the telephones connected to the amplifier system. In order that this condition may be attained the emf. across the resistance R_2 must be equal in magnitude and opposite in sign to that across R_1. The proper phase relationship is provided by the mode of connection. The satisfaction of the other condition may be identified with the expression:

$$I\,R_1 = I_p\,R_2 \tag{a}$$

But, by definition,

$$I_p = \rho\,E_g = \rho\,R\,I \tag{b}$$

which, substituted in (a), leads to

$$I\,R_1 = \rho\,R\,I\,R_2 \tag{c}$$

and

$$\rho = \frac{1}{R} \cdot \frac{R_1}{R_2}. \tag{d}$$

Also, since R is purposely set at 1,000 ohms, we have finally

$$\rho = .001\,\frac{R_1}{R_2} \quad \text{amps/volt} \tag{e}$$

which is a very simple relation and therefore one particularly suited to the rapid execution of a large number of measurements.

The terminals of the slide wire, R_1R_2 are connected to the first tube of an audio frequency vacuum tube amplifier in the manner shown. It is rather important that the biasing battery, C, be connected in series with the grid circuit in order that the resistance of this circuit shall be infinite and there will be no tendency for the currents in the slide wire to flow thru the grid circuit, which would happen if the grid were allowed to assume a positive potential at the operating point. Under these conditions, if care has been taken in the arrangement of the wiring,

the minima will be well defined. The resistance of the slide wire may well be of the order of 100 ohms so that the potential drops may be greater and a better indication may be had with a minimum of amplification. In the writer's set-up this slide wire consisted of the slide wire portion of the "Student's Potentiometer" manufactured by the Leeds-Northrup Company, which had a total resistance of 116 ohms, and which proved to be very well suited to the work having a scale graduated in 0.5 divisions from 0 to 100.

ADDENDUM II†

As an illustration of the great ease and simplicity with which problems relating to vacuum tube circuits may be handled with the aid of the fundamental plate circuit theorem and the conceptions of the voltage amplification factor and internal impedance, we may profitably consider the oscillation circuit depicted in Figure B, which has previously been treated by Hazeltine.[1]

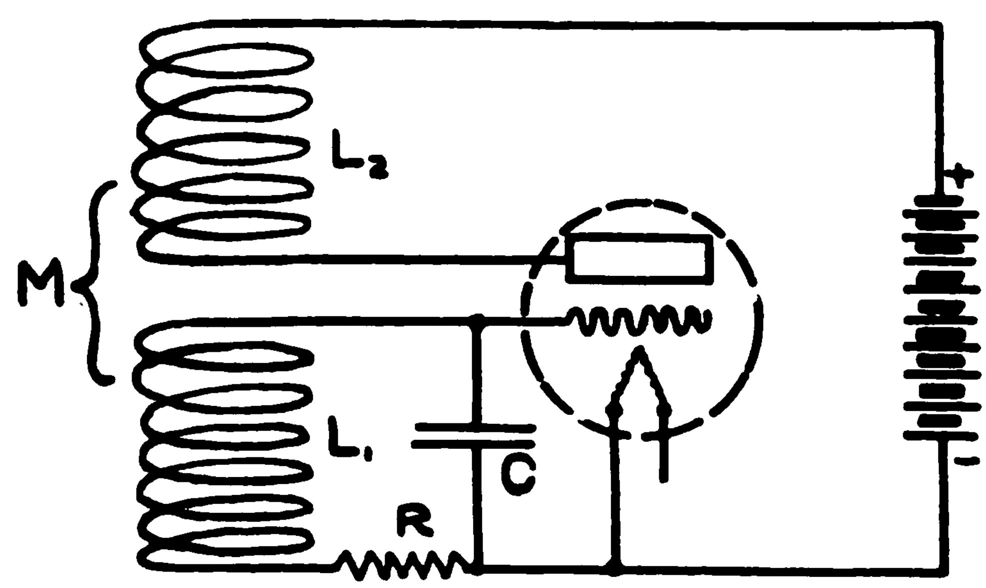

FIGURE B

This circuit is extensively used for receiving purposes in connection with the autodyne method of reception of continuous wave signals. It also enjoys extensive application in transmitting, but when used for this purpose, the entire characteristic curve is utilized, and the grid potential may vary over widely separated limits rendering the rigorous analysis of the operation difficult, if not impossible. In order to construct the differential equations for the potentials, it would be necessary first to formulate the dependence of the amplification factor and internal

† Received by the Editor, January 15, 1919.
[1] Hazeltine, previous citation, page 79.

impedance upon the instantaneous value of the grid potential, e_g. This could be done, possibly, with the aid of empirical expressions covering the variation of μ and R_o as experimentally determined, but it is doubtful if a simple integration of the equations so formed could be effected. For this reason and for the sake also of simplicity, in the present discussion it will be assumed that operation involves only a very small part of the characteristic surface, and that in the region considered the values of μ, and R_o are fixed. This is not so bold an approximation as it would at first seem, since the most favorable part of the static characteristic curve is at the point of inflection and in this region the amplification factor and internal impedance are practically constant. This is shown by the experimental curves displayed in the paper. After the oscillation has started, of course, the operating region expands until the slope of the curve decreases below the minimum possible value of ρ determined by the coefficients of the circuit. The problem consists in formulating the relation between the tube parameters and the circuital coefficients which will render sustained oscillation of the system possible. As mentioned above, this matter has been discussed by Hazeltine in a very valuable paper covering various types of oscillation circuits. In his discussion, the assumption is made that the plate current is a simple function of the grid voltage, the relation being $i_p = \rho\, e_g$. Any reaction existing between the two circuits is thereby not considered, a result being obtained, which, while very simple and useful, cannot be regarded as a complete definition of the criterion for the oscillation in actual systems. The following mode of treatment is more fundamental and the effect of reaction is completely determined.

Referring to Figure B, the circuital constants are denoted by the conventional symbols, the internal impedance of the plate circuit being inserted in series with the inductance, L_2 and an emf. of value $\mu\, e_g$, in accordance with the usual method of considering the plate circuit. The equations for the potentials may be constructed as follows:

$$L_1 \frac{di}{dt} + Ri + \frac{1}{C}\int i \cdot dt + M\frac{di_p}{dt} = 0 \tag{1}$$

$$L_2 \frac{di_p}{dt} + R_o i_p + M\frac{di}{dt} + \frac{\mu}{C}\int i \cdot dt = 0 \tag{2}$$

since

$$e_p = \mu\, e_g = \frac{\mu}{C}\int i \cdot dt.$$

We are interested primarily in the variable, i, representing the current in the oscillatory circuit $L_1 C$, which may be isolated by solving (1) for $\frac{d i_p}{dt}$ and substituting the value found in (2). After multiplying thru by $(-M)$, differentiating twice with respect to t, and collecting, we have:

$$(L_1 L_2 - M^2)\frac{d^3 i}{d t^3} + (R L_2 + R_o L_1)\frac{d^2 i}{dt} + \left(\frac{R R_o C + L_2 - \mu M}{C}\right)\frac{di}{dt} + \frac{R_o}{C} i = 0 \qquad (3)$$

which may be placed in the more convenient form:

$$\frac{d^3 i}{dt^3} + \beta\frac{d^2 i}{dt} + \gamma\frac{di}{dt} + \delta = 0 \qquad (4)$$

This is a linear differential equation of a familiar type with constant coefficients in the operating region considered of the form:

$$\beta = \frac{R L_2 + R_o L_1}{L_1 L_2 a}$$

$$\gamma = \frac{R_o R C + L_2 - \mu M}{L_1 L_2 a C} \qquad (5)$$

$$\delta = \frac{R_o}{L_1 L_2 a C}$$

where a is the coefficient of leakage of dimensions, $1-k^2$, and depends upon the coupling between the grid and plate circuits.

Proceeding as usual with an assumed integral of the form:

$$i = A\,\varepsilon^{a t} \qquad (6)$$

substitution in (4) yields the cubic auxiliary:

$$a^3 + \beta a^2 + \gamma a + \delta = 0 \qquad (7)$$

to be solved for the a's. The complete solution is:

$$i = A\,\varepsilon^{a_1 t} + B\,\varepsilon^{a_2 t} + C\,\varepsilon^{a_3 t} \qquad (8)$$

where A, B, and C are constants of integration to be evaluated, when amplitudes are of interest, from a knowledge of the initial configuration of the system. The solution of the cubic may be effected by Ferreo's method, but the expanded result being practically unmanageable, this method of attack may well be abandoned. We may obtain some light on the probable nature of the roots by considering the physical phenomena. In order that these may be real and periodic, as expected, it will be neces-

sary for one of the roots to be real and the others complex. On this basis, we may write:

$$
\begin{aligned}
a_1 &= a_1 \\
a_2 &= \Delta + j\,\omega \\
a_3 &= \Delta - j\,\omega
\end{aligned}
\tag{9}
$$

A fundamental theorem in the theory of equations permits us to write also:

$$
\begin{aligned}
a_1 + a_2 + a_3 &= -\beta \\
a_1\,a_2 + a_1\,a_3 + a_2\,a_3 &= \gamma \\
a_1\,a_2\,a_3 &= -\delta
\end{aligned}
\tag{10}
$$

which applied to the present case establishes the relations:

$$-a_1 - 2\,\Delta = \frac{R\,L_2 + R_o\,L_1}{L_1\,L_2\,a} \tag{11}$$

$$2\,\Delta\,a_1 + \Delta^2 + \omega^2 = \frac{R\,R_o\,C + L_2 - \mu\,M}{L_1\,L_2\,a\,C} \tag{12}$$

$$-a_1\,(\Delta^2 + \omega^2) = \frac{R_o}{L_1\,L_2\,a\,C} \tag{13}$$

To approximate the roots of (7) we will resort to a very useful artifice employed by Dr. Fulton Cutting[2] in connection with his excellent treatment of the transient phenomena in radio transformer circuits, which consists essentially in taking the ratio of the equations (12) and (13)

$$-\frac{1}{a_1}\left(1 + \frac{2\,\Delta\,a_1}{\Delta^2 + \omega^2}\right) = \frac{R\,R_o\,C + L_2 - \mu\,M}{R_o} \tag{14}$$

thus obtaining a factor in the left hand member which is quite ignorable in comparison with unity. This is particularly permissible in the present case since, at the radio frequencies we are dealing with, ω is large compared with the damping. Proceeding with this we obtain:

$$-a_1 = \frac{R_o}{R\,R_o\,C + L_2 - \mu\,M} \tag{15}$$

and from the relation (11) the more important damping term:

$$\Delta = \frac{R_o}{2\,(R\,R_o\,C + L_2 - \mu\,M)} - \frac{R\,L_2 + R_o\,L_1}{2\,L_1\,L_2\,a} \tag{16}$$

which are excellent approximations. The general solution may be written in the form:

$$i = A\,\varepsilon^{-a_1 t} + \varepsilon^{\Delta t}\,(B\,\varepsilon^{j\omega t} + C\,\varepsilon^{-j\omega t}) \tag{17}$$

---

[2] Fulton Cutting, Proceedings of The Institute of Radio Engineers, volume 4, page 160, 1916.

Physically, the first term in the above solution is of the nature of a transient and disappears when the oscillations have reached their equilibrium state. The second term is of particular interest since it represents the periodic phenomena under investigation. In order that the oscillations may be continuous and self sustained, it is necessary that the effective damping of the circuit, represented by Δ, shall be zero. The satisfaction of this condition leads to:

$$\frac{R_o}{R\,R_oC+L_2-\mu M}=\frac{RL_2+R_oL_1}{L_1L_2a} \tag{18}$$

as an expression relating the coefficients of the circuit and the parameters of the tube for the condition of sustained oscillation. This may be solved for ρ giving:

$$\rho=\frac{\mu}{R_o}=\frac{RC}{M}+\frac{L_2}{M\,R_o}-\frac{L_1L_2a}{M\,(RL_2+R_oL_1)} \tag{19}$$

It is interesting to note that when the plate circuit inductance L_2 is ignored, and the induction effect into the grid circuit being retained thru the concept of a fictitious mutual inductance, this degenerates into:

$$\rho=\frac{C\,R}{M} \tag{20}$$

which is Hazeltine's result.

In most practical circuits, the term RL_2 may be ignored in comparison to R_oL_1 so that equation (19) may be simplified to:

$$\rho=\frac{C\,R}{M}+\frac{M}{L_1R_o} \tag{21}$$

A graphical representation of this expression is shown in Figure C. In constructing this curve the following circuital constants were assumed:

$$L_1=60\ \mu\text{h.}$$
$$L_2=180\ \mu\text{h.}$$
$$C=0.0002\ \mu\text{f.}$$
$$R=20\ \Omega$$
$$R_o=15{,}000\ \Omega$$

The component curves are shown by dashed lines, their dimensions being designated. The hyperbola is identical with the result obtained by Hazeltine, while the straight part increasing with M represents the effect of the reaction between the circuits. It is evident, from physical considerations, that an optimum value of M exists which is most favorable to oscilla-

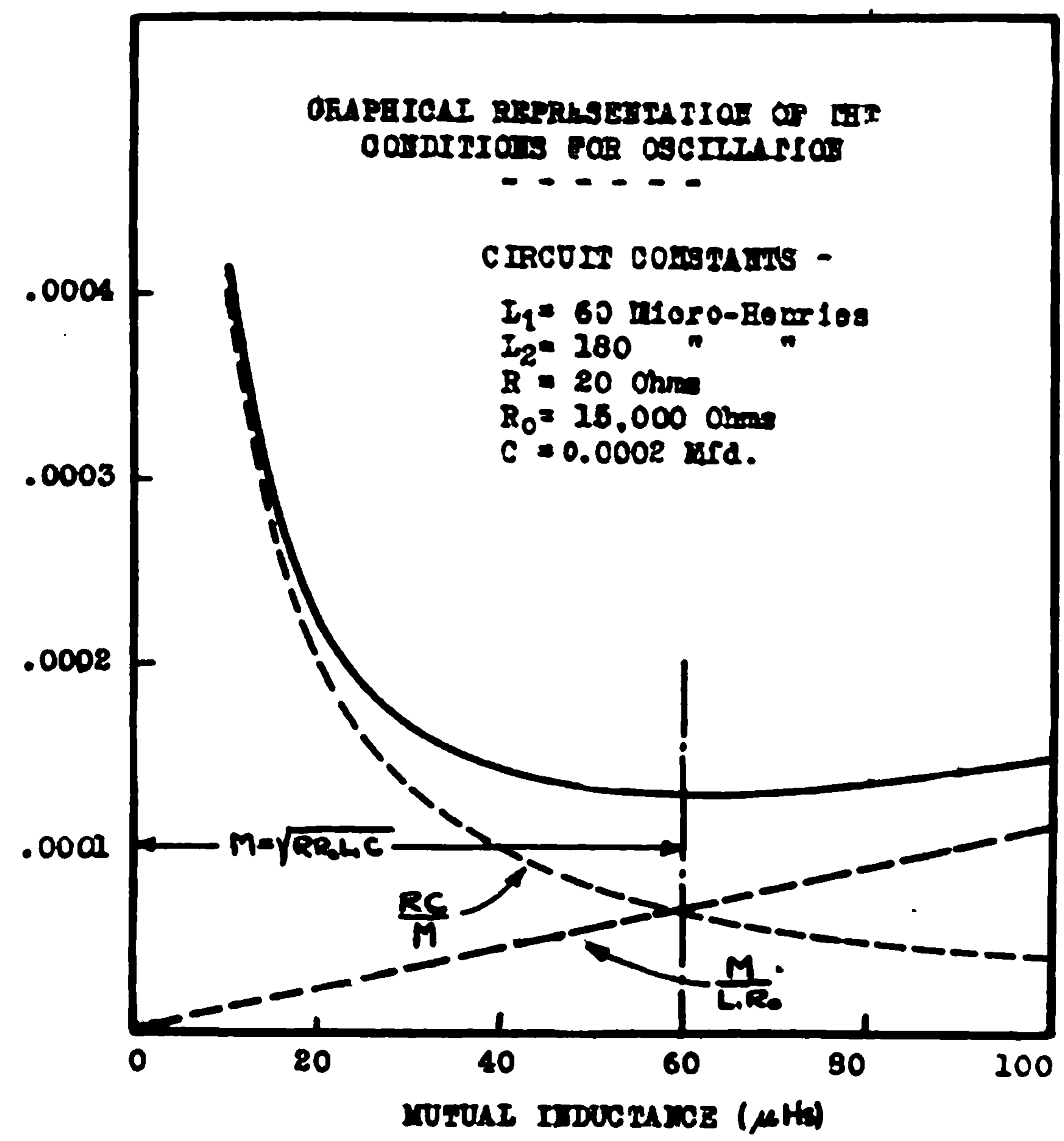

FIGURE C

tion. Differentiating (21) with respect to M and equating to zero we find this to be:

$$M_{opt} = \sqrt{R\, R_o L_1 C} \tag{22}$$

If it is assumed that this value of M is selected, and substituted in equation (21) we find that the minimum ρ is:

$$\rho = 2\sqrt{\frac{C\,R}{L_1\,R_o}} \tag{23}$$

or is just double its value in the absence of circuital reaction. Equation (22) may be re-written in terms of the wave length as:

$$M_{opt} = \frac{\lambda}{2\pi v}\sqrt{R\,R_o} \tag{24}$$

In practical systems, a few trial calculations will show that the point of optimum mutual inductance is seldom found, the circuit used in the above curves being somewhat of an ultimate limit. If the inductance in the plate circuit is increased so as to increase the coupling, an optimum coupling could be obtained with higher wave lengths and higher resistances in the oscillatory circuit, were it not for the fact that this brings us to the consideration of the inherent capacity of the plate circuit acting across the plate inductance. Under these conditions the plate

and grid circuits may be more or less in resonance and the above equations are not applicable. This does not decrease in any way the practical value of (21) since the reaction, no matter how small, is always effective in reducing the tendency toward oscillation.

The above equations may be further utilized in the study of the regenerative operation of the vacuum tube, by impressing emf.'s of suitable form upon the input circuit of the tube. The above roots are applicable in the formulation of the free solution, the forced solution being found in the usual way. This investigation will be undertaken in another paper.

SUMMARY OF ADDENDUMS: In a first addendum, the author shows a later method of measuring directly the mutual conductance of tubes, which method is free from certain possible objections of the earlier procedure.

In a second addendum, an autodyne oscillator with inductive coupling between grid and plate circuits is analytically considered. The condition for sustained oscillation is derived, a numerical illustration given, and the value of the optimum mutual inductance between grid and plate circuits for oscillation is obtained.

DISCUSSION

H. J. van der Bijl (by letter): Without entering into a full discussion of Mr. Ballantine's paper, I wish to call attention to a few outstanding points.

The method he gives for measuring the mutual conductance is very interesting and obviously allows of a simple and rapid determination of this important quantity. It might, however, be well to point out that the method, as it stands, is only applicable to cases in which the output impedance of the tube is large compared with that of the coil connected in the output circuit. As long as this is so, the alternating current, i_p, in the plate circuit can (for small impressed oscillations) be given by the simple expression used by Mr. Ballantine, namely: $i_p = \rho\, e_g$ where $\rho = \frac{\mu}{R_o}$, the mutual conductance, and e_g is the a. c. input voltage. If, on the other hand, the external impedance, Z, into which the tube works is not negligibly small, the current is given by

$$i_p = \frac{\mu\, e_g}{R_o + Z}$$

which follows directly from the amplification equations I gave in my paper cited. (μ as used here is the same as μ_o in my equations.) This then gives

$$\frac{R_o}{\mu} = \frac{e_g}{i_p} - \frac{Z}{\mu}$$

where $\frac{R_o}{\mu}$ is the inverse slope of the static characteristic and $\frac{e_g}{i_p}$ the inverse slope of the dynamic characteristic, or writing ρ for the true mutual conductance of the tube and ρ' for that of the tube and circuit:

$$\frac{1}{\rho} = \frac{1}{\rho'} - \frac{Z}{\mu}.$$

Mr. Ballantine does not state the impedance of the coil used in his experiments, but it was undoubtedly small in comparison with that of the tube which, at the voltages used, had a rather high impedance. The second term of the above equation can, however, become very disturbing when using low impedance tubes, and tubes having an impedance of a few thousand and even a few hundred ohms are not uncommon. In any case the external impedance Z must be made negligibly small, otherwise the necessity of taking it into consideration, thus

compelling the use of the above equation, instead of the simple one given by Mr. Ballantine, destroys the usefulness of his method which aims at being quick and simple.

This consideration shows that the mutual conductance depends on the constants of the circuit in which the tube is operated, having a limiting value, $\frac{\mu}{R_o}$, which obtains when the tube works into a negligibly small impedance, and which can be characterized as the mutual conductance of the tube itself. The reason for this becomes readily apparent when considering that the external impedance tends to straighten out the characteristic. Even if the external impedance were a pure reactance, the dynamic characteristic would tend to straighten out and have a smaller slope than the static characteristic which would not be affected by the reactance. If, for example, the current be increased by increasing the grid potential, the voltage between filament and plate decreases, due to the increased voltage drop in the coil. In other words, both E_g and E_p in the characteristic equation are variables, E_p being always 180° out of phase with E_g. Mr. Ballantine's method could, of course, be used as it stands to determine the particular value the mutual conductance would have in any particular circuit, (say the circuit in which the tube is to be operated in practice) by choosing an impedance in the test circuit equal to that to be used in practice. This would, however, not be so simple as computing the mutual conductance of the tube itself from μ and R_o and then determining its value for any circuit with the help of the above equation. I think that a determination of μ and R_o is more important because a direct determination of the mutual conductance tells nothing about μ or R_o, the values of which must be known in order to determine the operating range of the tube, the impedance to be used in the output, and so on. On the other hand, a separate determination of μ and R_o allows the computation of the mutual conductance for any type of circuit. It is very important to distinguish between the static characteristic of the tube, which is governed by the fundamental characteristic equation, and the dynamic characteristic of the tube *and circuit*, which is governed by the amplification equations I gave in my paper in this issue of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS.

The other important point is the limits of operation of the tube. For example, referring to Figure 6 of his paper, which gives the amplification constant μ as a function of the grid

voltage, Mr. Ballantine states that "The salient feature of these curves lies in the fact that the amplification factor is not constant, as has been intimated from measurements with direct currents, by Dr. van der Bijl, but depends greatly upon the grid potential in the region in which operation takes place." The amplification constant is a function only of the geometry of the tube, and must therefore be constant. When it is measured by the dynamic null method it will be found to be constant provided the tube is operated within the limits that I specified. These limits apply only to the grid voltage. (The equations are not by any means limited to small plate current.) As soon as the grid becomes sufficiently positive to take current the effective grid voltage is reduced, and unless this reduction is taken into consideration μ will come out too small, as can easily be seen by considering the circuit for the dynamic measurement of μ. It is obvious that this effect increases as the plate voltage is decreased because a larger plate voltage would have a greater influence in pulling the electrons thru the grid.

For negative values of the grid potential the limitation is imposed mainly by $\frac{E_p}{\mu}$, and will be more effective the lower the value of E_p. These considerations, I think, explain the variation in μ with grid potential observed by Mr. Ballantine. The tube with which he obtained his curves has a very small operating range of grid voltage which, of course, depends on the plate voltage used. In a tube of these general constants the value of ε in my characteristic equation becomes an important quantity, as it always is when $\frac{E_p}{\mu}+E_g$ becomes small. In fact ε depends to a great extent on the nature and surface condition of the electrodes. If it is positive the μ, E_g-curve will drop more readily for positive values of the grid potential, and if negative the effect will be on the negative side. I have obtained curves like these with a Western Electric Company tube of the telephone repeater type and found that with 150 volts on the plate, μ which had a value of 5.0 remained constant to within 10 per cent over a range of grid voltage from -20 to $+20$. This is because this type of tube has a greater operating range and is, therefore, not so easily overtaxed.

L. A. Hazeltine*: 1. MUTUAL CONDUCTANCE. The writer wishes to congratulate Mr. Ballantine on the clever arrangement (Figure 2) that he has devised for directly measuring the mutual conductance of an amplifier bulb. This method readily gives the mutual conductance for any combination of direct-current adjustment (grid potential, plate potential and heating current), as shown by Mr. Ballantine's curves. It may be noted, however, that mutual conductance in general depends also on certain conditions of the alternating-current circuits, as follows:

(a) *On the frequency.* It is conceivable that the mutual conductance obtained by a dynamic method might differ from the slope of the static characteristic curve obtained with direct current; and it is also conceivable that values at radio frequency might differ from those at audio frequency. While differences undoubtedly occur between the static and dynamie characteristics in bulbs showing positive ionization, no such frequency effects are to be expected in high-vacuum bulbs, nor have any been found, so far as the writer is aware. If desired, however, Mr. Ballantine's method may readily be extended to radio frequencies by substituting for the simple buzzer a buzzer-excited oscillating source and for the audio-frequency balancing transformer a radio-frequency transformer together with a detector. The method may also be extended to direct-current operation by substituting two resistances and a galvanometer for the balancing transformer and telephone; but some modifications are necessary on account of the normal direct current of the plate.

(b) *On the amplitude of oscillation.* With a low impressed alternating voltage, the variations of the grid voltage and plate current will be over such a short arc of the characteristic curve that this virtually coincides with its tangent. The slope of the latter therefore represents the mutual conductance for small amplitudes of oscillation. For higher impressed voltages the curvature of the characteristic will become appreciable, and the mutual conductance will then be rather crudely represented by the slope of the secant line connecting the extreme points of the oscillation[1]. If Mr. Ballantine's method is to be used for large amplitude of oscillation, the effect of wave distortion must be

*Revised and amplified from the oral discussion at the Institute meeting, New York, December 11, 1918. Received by the Editor, January 9, 1919.

[1]For the exact definition of the mutual conductance with large amplitudes of oscillation, see the writer's paper, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 64, foot-note.

eliminated, either by sharply tuning the telephone circuit to the supply frequency or by substituting for it a sharply tuned vibration galvanometer.

(c) *On the presence of an alternating plate voltage.* If the plate potential is varied proportionally to the grid potential but in an opposite sense (as must happen in any use of the bulb in which power is given out from its plate circuit), the changes in plate current will be smaller than if the plate potential remained fixed. The "effective mutual conductance" (see Article 3 of this discussion) is therefore lowered by the presence of an alternating plate voltage. Mr. Ballantine's method can be extended to give this effective mutual conductance by including part of the resistance R (Figure 2) in the plate circuit.

In calling attention to the possible extensions of Mr. Ballantine's method, the writer appreciates that the low-amplitude value of mutual conductance, unmodified by the presence of an alternating plate voltage, is the most fundamental and most useful value, and that the simple method of the paper is far more likely to be of practical value than any of the extensions.

In regard to the symbol for mutual conductance, the writer still prefers g to ρ, considering it very desirable to denote quantities of the same physical dimensions by the same symbol. This has been found particularly true for mutual conductance, as it often occurs in formulas similar to those containing other conductances (for example, compare equations (26) and (27) below). In cases where more than one conductance enters in the same equation, different subscripts or other distinguishing marks may be used. The writer suggests denoting mutual conductance by g_m in such cases, and will use this notation in what follows.

2. PLATE RESISTANCE. The term "internal impedance" is open to two objections: (a) the bulb has *two* "internal" circuits, that of the grid as well as that of the plate; and (b) the impedance has the nature of a pure *resistance*, as is assumed in all proposed methods for measuring it, and so for clearness' sake should be called a resistance. The proper term for this constant would seem to be "plate resistance" and the appropriate symbol r_p. The use of the reciprocal, the plate conductance g_p, is sometimes more convenient.

In this connection it might be well to compare the two methods of viewing the plate circuit, considered as a source of alternating-current power. According to the first method, the alternating plate current I_p is made up of two terms, one due to

the alternating grid voltage E_g and the other due to the alternating plate voltage E_p:

$$I_p = E_g\, g_m - E_p\, g_p, \tag{1}$$ [2]

where g_m is the mutual conductance and g_p is the plate conductance. Interpreting this equation, we may say that *the plate circuit acts as a generator whose total current generated is $E_g\, g_m$, of which a part $E_p\, g_p$ is lost in a shunt path, or "leak," of conductance g_p.* To lead to the second method of viewing the action, we may solve equation (1) for E_p:

$$E_p = \frac{g_m}{g_p} E_g - \frac{I_p}{g_p}; \tag{2}$$

or

$$E_p = \mu E_g - I_p\, r_p, \tag{3}$$

where

$$\mu = \frac{g_m}{g_p} \tag{4}$$

is the amplification constant, and

$$r_p = \frac{1}{g_p} \tag{5}$$

is the plate resistance. Interpreting equation (3), we may say that *the plate circuit acts as a generator the generated or internal voltage of which is μE_g and the internal resistance of which is r_p.*[3]

These two methods are equivalent and therefore equally correct. That represented by equation (1) probably corresponds most closely to the physical action, and is most useful in treating oscillating circuits. That represented by equation (3) has the advantage that one of the constants (μ) is nearly independent of the direct-current adjustments of the bulb, and is most useful in treating amplifiers and detectors.

3. Use of "Effective Mutual Conductance." In many calculations of thermionic oscillators it is found convenient

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[2] This equation was given by the writer, Proceedings of The Institute of Radio Engineers, volume 6, page 67. It is the alternating-current equivalent of Vallauri's equation, which in corresponding notation is

$$i_p = e_g\, g_m + e_p\, g_p + c.$$

Equations of this form seems to have been first employed by Latour, "Electrician," (London), December 1, 1916 (in discussing amplification), and by Bethenod, "La Lumière Electrique," December 16, 1916 (in discussing oscillation).

[3] The equivalent of equation (3), its derivation and its interpretation, were given by Miller, Proceedings of The Institute of Radio Engineers, volume 6, page 143.

to deal with a single constant, rather than with two independent constants (g_m and g_p, or μ and r_p). That is, the "effective mutual conductance" g_m' is employed instead of the "normal mutual conductance" g_m, and is defined as the quotient of the (r.m.s.) plate current I_p by the (r.m.s.) grid voltage E_g, taking into account the presence of such alternating plate voltage as may exist under operating conditions:

Effective mutual conductance, $$g_m' = \frac{I_p}{E_g}. \tag{6}$$

Substituting (1),

$$g_m' = g_m\left(1 - \frac{g_p}{g_m} \cdot \frac{E_p}{E_g}\right); \tag{7}$$

or

$$g_m' = g_m\left(1 - \frac{n}{\mu}\right), \tag{8}$$

where

$$n = \frac{E_p}{E_g} \tag{9}$$

is the ratio of plate voltage to grid voltage as determined by the circuit conditions. The expression $\left(1 - \frac{n}{\mu}\right)$ thus enters as a correction factor, which frequently differs from unity by a percentage less than the uncertainties in the values of g_m with various bulbs or various adjustments, and so may then be disregarded. On the whole, the effective mutual conductance g_m' is of more service in the calculation of oscillating-current circuits than the normal value; it is the value given by the equations for g in the writer's Institute of Radio Engineers paper previously referred to, some of which are included in Figure 2 herewith.

It is interesting to observe that the effective mutual conductance g_m' for any chosen value of n may be determined with a single setting of Dr. Miller's apparatus, as follows. In Figure 1 herewith[4] the ratio r_2/r_1 is set at the desired value of n, and R is adjusted for a balance. We then have

$$g_m' = \frac{I_p}{E_g} = \frac{E_p/R}{E_g} = \frac{n}{R}. \tag{10}$$

Of course, if n is very small or very near μ [which it cannot exceed according to (8)], the precision will be low.

[4] Figure 1 of Dr. Miller's paper, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, page 144, 1918.

Certain cases arise where it is necessary to employ the correction factor $\left(1-\frac{n}{\mu}\right)$ of equation (8), or to use a value of g_m' differing considerably from g_m. Three of these will be considered here.

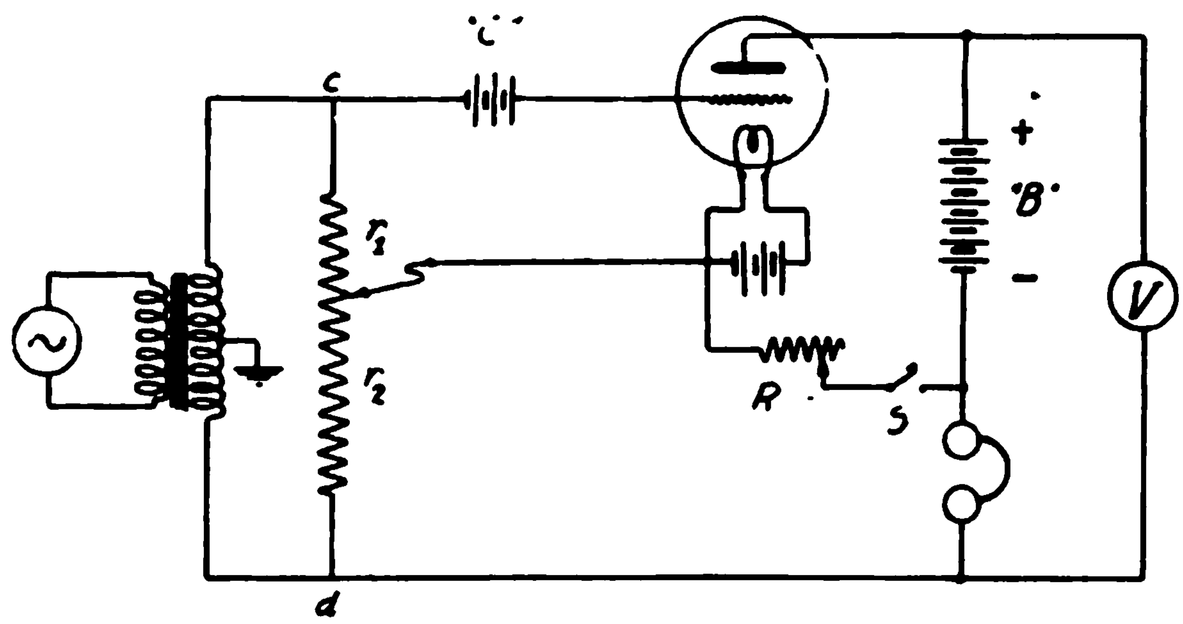

FIGURE 1

(*a*) *Optimum Mutual Inductance.* In Figure 2b herewith (Figure 3b of the writer's paper) the effective mutual conductance (denoted simply as g), is

$$g_m' = \frac{C\,r}{M}; \tag{11}$$

so the normal mutual conductance by (8) is

$$g_m = \frac{C\,r}{M\left(1-\frac{n}{\mu}\right)} = \frac{C\,r}{M\left(1-\frac{1}{\mu}\cdot\frac{M}{L}\right)}. \tag{12}$$

Considering M as variable, it can readily be shown that g_m will be a minimum (or the oscillation will be the strongest) when

$$\frac{M}{L} = \frac{\mu}{2}. \tag{13}$$

Substituting in (12), the value of g_m is found to be $\frac{2\,C\,r}{M}$ or just *twice* the effective mutual conductance in (11). In other words, to maintain a strong oscillation an optimum value of mutual inductance occurs, as given by (13); and for this value the effective mutual conductance has fallen to one half the normal value.[5]

[5]The existence of an optimum mutual inductance was first brought to the writer's attention by Mr. Ballantine, who has derived an expression for its value differing in form from equation (13) but equivalent thereto. (This is found on page 160 of Addendum 2 above.—Editor.)

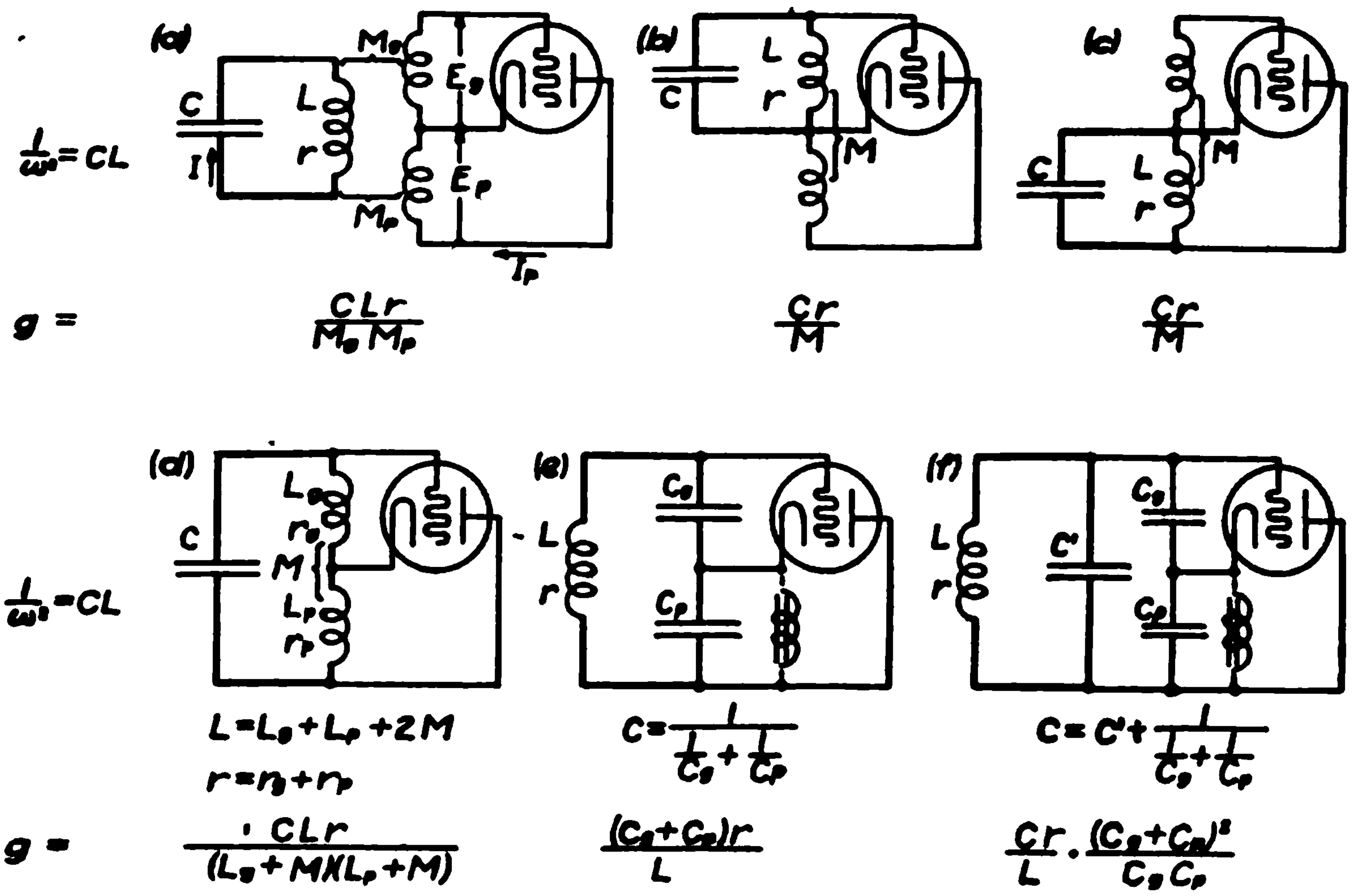

FIGURE 2

(*b*) *Optimum Voltage Division.* In Figure 2d herewith the effective mutual conductance is given as

$$g_m' = \frac{CLr}{(L_g+M)(L_p+M)} = \frac{Cr}{L} \cdot \frac{(E_g+E_p)^2}{E_g E_p} \tag{14}$$

$$= \frac{Cr}{L} \cdot \frac{(1+n)^2}{n}. \tag{15}$$

This is a minimum when $n=1$; so it has sometimes been thought that in circuits of this form the oscillation is the strongest when the total voltage is divided equally between the plate and the grid. However, introducing the correction factor, we have

$$g_m = \frac{Cr}{L} \cdot \frac{(1+n)^2}{n\left(1-\frac{n}{\mu}\right)}, \tag{16}$$

which is a minimum for

$$n = \frac{\mu}{\mu+2}. \tag{17}$$

That is, the oscillation will be the strongest when the total voltage is divided between the plate and the grid in the ratio $\mu/(\mu+2)$, or about 0.8 for values of μ usually found in detector bulbs. For this case the correction factor becomes

$$1-\frac{n}{\mu} = \frac{\mu+1}{\mu+2}, \tag{18}$$

which is usually in the neighborhood of 0.9.

The above deduction applies also to Figure 2e with fair accuracy, but to Figure 2f only very roughly, especially when C_g and C_p are small compared with C' and when $C\,r/L$ is high. Exact treatment of the last case requires the complex method, for the reason that on account of the grid and plate conductances the voltages across C_g and C_p may be considerably out of phase with one another, contrary to the assumptions made in the "loss method" used in deriving the formulas of Figure 2. (Both theoretical and experimental results show optimum values of n varying over a considerable range as the circuit constants are varied, and sometimes more than one optimum value occurs in a given circuit.)

(c) *Maximum Output.* The circuit of Figure 2a herewith is frequently used in bulb transmitters, for which we wish to choose the circuit constants to secure the greatest output. The effective mutual conductance is given as

$$g_m' = \frac{C\,L\,r}{M_g\,M_p}; \tag{19}$$

and the voltage ratio is evidently

$$n = \frac{M_p}{M_g}. \tag{20}$$

Solving for M_g and M_p,

$$M_g = \sqrt{\frac{C\,L\,r}{g_m'\,n}} \qquad \text{and} \qquad M_p = \sqrt{\frac{C\,L\,r\,n}{g_m'\,n}}. \tag{21}$$

Now the output of the bulb depends only on g_m' and n (as far as the radio-frequency circuit is concerned); so if their optimum values can be determined M_g and M_p can be at once computed by (21).

The writer has previously indicated[6] how the values of n and g_m' for greatest output could be determined by trial from the characteristic curves of the bulb. A more convenient method would probably be by direct experiment, using any of the circuits of Figure 2. Thus if Figure 2c is employed, we may insert a radio-frequency ammeter in the circuit (C, L) and vary M and C (or M and r) until a maximum output is obtained, varying also the direct-current quantities within permissible limits to determine their most desirable adjustments. When the best arrangement has been found, the desired values are

$$g_m' = \frac{C\,r}{M} \tag{22}$$

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[6] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 6, pages 94 and 95.

and

$$n = \frac{L}{M}. \tag{23}$$

4. **Detector Constants.** The constants which Mr. Ballantine denotes by D and σ and calls variously "detection," "detecting constant," "detector constant," "detecting action" and "detection factor," are open to criticism. Neither is defined directly in terms of the quantities in which the user of a detector is interested; and—more unfortunately still—these two constants, "detection (without grid condenser)" and "detection (grid condenser)," are not of the same physical dimensions and so are not numerically comparable with one another, as one might have expected. Several months ago the writer used the term "detector constant" in a private report to represent *the change in plate current per volt square (r.m.s.) impressed on the grid.* This definition leads to the following formulas:

$$\text{Detector constant without grid condenser} = \frac{1}{2} \cdot \frac{d^2 i_p}{d e_g^2}, \text{ and} \tag{24}$$

$$\text{Detector constant with grid condenser} = \frac{1}{2} \cdot \frac{d i_p}{d e_g} \cdot \frac{\dfrac{d^2 i_g}{d e_g^2}}{\dfrac{d i_g}{d e_g}} \tag{25}$$

(where small letters are used, instead of the corresponding capitals of Mr. Ballantine, to represent the instantaneous currents and voltages, as is customary in alternating-current circuit equations). The former [equation (24)] agrees with Mr. Ballantine's D except for the factor $\frac{1}{2}$; but the latter [equation (25)] differs from his σ not only by the factor $\frac{1}{2}$, but also by the factor $\frac{d i_g}{d e_g}$ in the denominator. Mr. Ballantine's omission of this factor is distinctly in error; for if two bulbs are alike in all other respects, that having the lower value of $\frac{d i_g}{d e_g}$ will be a correspondingly better detector—quite irrespective of any of the incidental features of operation which Mr. Ballantine mentions in connection with the definition of this constant.

Very recently the writer has used the term *rectification constant* to represent *the change in the grid generated (d.c.) voltage per volt square (r.m.s.) impressed* and considers this a more fun-

damental and more useful constant than the above "detector constant." The formulas for this constant without and with a grid or stopping condenser are respectively,

Mutual rectification constant, $$\nu_m = \frac{1}{2} \cdot \frac{\dfrac{d^2 i_p}{d e_g^2}}{\dfrac{d i_p}{d e_g}} = \frac{1}{2 g_m} \cdot \frac{d g_m}{d e_g}, \text{ and} \tag{26}$$

Grid rectification constant, $$\nu_g = \frac{1}{2} \cdot \frac{\dfrac{d^2 i_g}{d e_g^2}}{\dfrac{d i_g}{d e_g}} = \frac{1}{2 g_g} \cdot \frac{d g_g}{d e_g}. \tag{27}$$

The theory of rectification in various forms of detectors and methods of directly measuring the rectification constants are included in a paper now in preparation.

The connections of Figure 4 for determining σ are also open to criticism, for they fix the grid potential arbitrarily. Actually the potential of the grid is fixed by the grid leak resistance (added or inherent), whose value has a very marked influence on the value of σ. This is because the entire direct current of the grid circuit must return thru the leak; so the operating point on the grid characteristic curve will be that for which the quotient of grid potential by grid current is equal to the resistance of the leak. In the report referred to above, the writer used connections similar in principle to Figure 4, but elaborated so as to determine the actual operating potential of the grid and to measure the grid conductance at and near this potential. This method has been used by the writer and by others with satisfactory results.

In regard to the practical value of such "detector constants" or "rectification constants" as measures of a bulb's effectiveness in radio reception, the following may be stated. They do not tell the whole story; for the bulb affects the circuit in more than one way. These constants do, however, give definite measures of certain important features of the bulb; and, together with the other necessary bulb data, they permit the exact calculations of the over-all effectiveness of a bulb used in connection with any given radio receiver. The other independent bulb constants which affect its detecting action are the amplification constant, the grid conductance, and the plate conductance (or resistance). If, for example, we compare the detector action of a bulb without and with a stopping condenser, we find the rectification constant several times higher *with* the stopping condenser. On the other

hand, when we calculate the effect of the high grid conductance inherent to operation with a stopping condenser, we find that it greatly reduces the radio-frequency voltage received by the grid and so may more than compensate for the better rectification. This agrees with practical experience, especially with grid circuits having a high quotient of self-inductance by capacity; for in such cases the grid loss may be several times all other radio-frequency losses put together, causing a marked diminution in impressed grid voltage.

V. Bush: Mr. Ballantine's very interesting paper is of particular service in gathering together and carefully defining the various tube constants which have been used by many writers. It is also a step toward a consistent accepted nomenclature of the subject, which will prove very necessary if we are to avoid confusion.

It would be an additional help if Mr. Ballantine could throw some light upon the following point: It seems to me that there should be certain standard conditions under which the constants of a tube are defined. We give the regulation of a generator, for instance, as the rise in terminal voltage from full load to no load under normal speed and field excitation. Could we not similarly more definitely define the internal impedance of a tube as the slope of the E_p, I_p curve under analogous standard conditions as regards plate voltage, filament current, and grid potential? Is it not true that the so-called constants of the tube reviewed in the paper are as yet not constants at all; but that we must give their values by means of a large number of curves for various conditions in order to describe completely the electrical operation of a given tube? Much of this is undoubtedly due to the versatility of the three element tube, so that many statements are needed to describe its action completely. Thus, for instance, we have no definite constant as yet for specifying output capacity. It is to be hoped however that as the art develops some definite set of constants and curves may be settled upon, as few in number as possible, which will describe briefly and completely the electrical behavior of a given tube under any set of conditions usually met with in practice, without adding a great deal of information which will not be needed.

With the ordinary generator we can of course go much further. There are various mechanical constants such as peripheral speed the constants of the iron, numbers of turns, and so on; and when these are known we can by means of well known formulas ex-

press the electrical constants in terms of them. That is, we can design consistently in order to obtain a generator for any desired purpose. It is of course too early to hope for such design formulas for thermionic tubes, for they are at present largely empirical.

Mr. Ballantine's paper will assist, however, toward the time when we will order a tube for a given service in exactly the same way that we now specify the behavior of a generator.

Edward Bennett (by letter)*: The use of uniform names and symbols for the constants of tri-electrode amplifiers is a matter of such importance that I venture to supplement the compilation of terms and symbols which the paper contains by a statement of the nomenclature which has been found very useful in an extended investigation of amplifiers under conditions in which the grid current cannot be neglected.

This nomenclature is as follows:

The ratio of the grid alternating current to the grid alternating potential (the plate potential being kept constant) is called the *grid conductance*, G_g. If ΔI_g and ΔE_g represent corresponding increments in the grid current and grid potential as read off from the continuous potential characteristics, then

$$\frac{\Delta I_g}{\Delta E_g} \text{ is represented by } G_g \tag{1}$$

The ratio of the plate alternating current to the plate alternating potential (the grid potential being kept constant) is called the *plate conductance*, G_p.

$$\frac{\Delta I_p}{\Delta E_p} \text{ is represented by } G_p \tag{2}$$

The ratio of the plate alternating current to the grid alternating potential (the plate potential being kept constant) is called the *controlled conductance of the plate by the grid*, or briefly the *controlled plate conductance*, G_{cp}.

$$\frac{\Delta I_p}{\Delta E_g} \text{ is represented by } G_{cp} \tag{3}$$

The ratio of the grid alternating current to the plate alternating potential (the grid potential being kept constant) is called the *controlled conductance of the grid by the plate*, or briefly the *controlled grid conductance*, G_{cg}.

$$\frac{\Delta I_g}{\Delta E_p} \text{ is represented by } G_{cg} \tag{4}$$

* Received by the Editor, January 16, 1919.

For the use of the term "conductance" in connection with the constants of amplifiers we are indebted to Professor Hazeltine. The quantity which in the above notation is termed the *controlled plate conductance* is by Professor Hazeltine termed the *mutual conductance*. It seems to me that the term mutual is open to objection in that the effect is not a mutual effect in the same sense as in mutual inductance or mutual elastance.

As an illustration of the utility of the constants defined above, consider their application to the simple case of non-regenerative power amplification in the circuit of Figure 1. In this figure A represents the source delivering the power which is to be amplified. The resistance, R, in the plate circuit represents the element to which the amplified power is to be delivered.

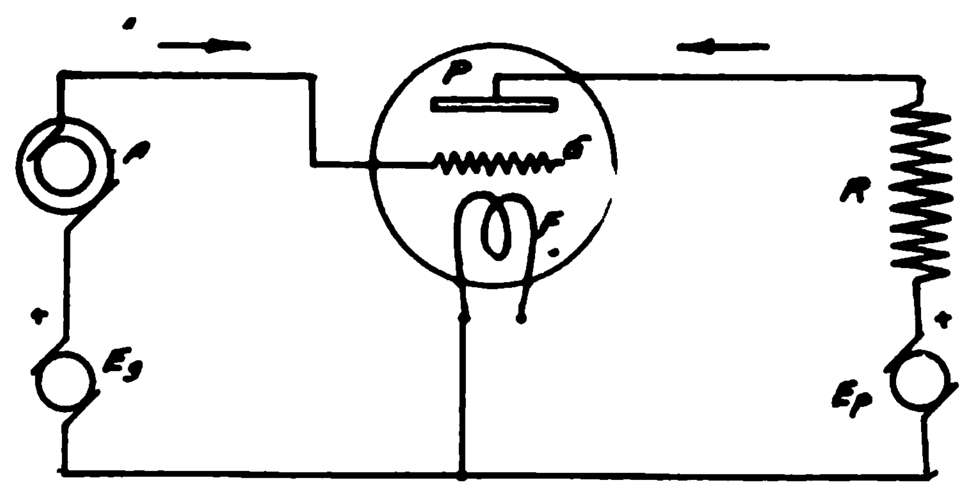

FIGURE 1

Let E represent the r.m.s. value of the alternating voltage of the source A supplying the power which is to be amplified.

Assuming for the moment that no variation occurs in the voltage between the plate P and filament F, the voltage E of the source impressed in the grid circuit will cause the following currents to flow:

$$I_{p1}=G_{cp}E$$

$$I_{g1}=G_gE$$

The passage of the alternating current of the value I_p thru the resistance R will, however, cause the plate voltage to vary by the amount,

$$E_1=-I_p R$$

This variation of the plate voltage will give rise to the following plate and grid currents,

$$I_{p2}=(-I_p R)G_p$$

$$I_{g2}=(-I_p R)G_{cg}$$

Whence the resultant plate and grid currents are as follows,

$$I_p\,(=I_{p1}+I_{p2})=G_{cp}E-I_p\,RG_p$$

or

$$I_p=\frac{G_{cp}E}{1+G_p\,R} \tag{5}$$

$$I_g\,(=I_{g1}+I_{g2})=G_gE-I_p\,RG_{cg}$$

or

$$I_g=G_gE-\frac{RG_{cp}G_{cg}E}{1+G_p\,R} \tag{6}$$

The power expended in the resistance is $I_p{}^2\,R=\left(\frac{EG_{cp}}{1+RG_p}\right)^2R$

The power delivered by the source A is

$$E\,I_g=E^2G_g-\frac{RG_{cp}G_{cg}}{1+RG_p}E^2$$

The power amplification =

$$\frac{I_p{}^2\,R}{E\,I_g}=\frac{G_{cp}^2\,R}{(1+G_p\,R)\,(G_g+G_gG_p\,R-G_{cp}G_{cg}\,R)} \tag{7}$$

The value which the resistance R in the plate circuit must have in order to lead to the maximum possible amplification of the power may be determined by taking the derivative of the amplification with respect to R, equating the derivative to zero, and solving the resulting equation for the value of R.

The value of R for maximum power amplification is found to be,

$$R_m=\frac{1}{G_p\sqrt{1-\dfrac{G_{cp}G_{cg}}{G_p\,G_g}}} \tag{8}$$

Substituting this value of R in equation (3), the expression for the maximum amplification is found to be,

Maximum power amplification =

$$\frac{G_{cp}{}^2}{G_gG_p\left[1+\sqrt{1-\dfrac{G_{cp}G_{cg}}{G_p\,G_g}}\right]^2} \tag{9}$$

In most tri-electrode devices, the fraction $\left(\frac{G_{cp}}{G_p}\right)\left(\frac{G_{cg}}{G_g}\right)$ is small in comparison with unity because of the small value of G_{cg}, or of the second fraction in comparison with the first.

Under these conditions the following approximate expressions may be written for maximum amplification.

$$R_m \text{ should equal } \frac{1}{G_p} \tag{8a}$$

$$\text{Maximum power amplification} = \frac{G_{cp}{}^2}{4G_g G_p} \tag{9a}$$

$$\text{The corresponding voltage amplification} = \frac{G_{cp}}{2G_p} \tag{6a}$$

The maximum voltage amplification, namely, $\frac{G_{cp}}{G_p}$, occurs only when the resistance R is made infinitely great; in this case the power amplification is 0.

Stuart Ballantine (by letter):* Dr. van der Bijl's remarks relative to the presence of inductance in the plate circuit of the vacuum tube under measurement with its concomitant effect on the value of the mutual conductance obtained, reflect a popular view concerning the accuracy of the method which is quite erroneous. It is true that the output circuit does contain the winding P_2 of the balancing transformer, which, in the absence of the supply voltage, e, will cause the current in this circuit to assume another value and lag behind the plate voltage by a small angle which depends upon the internal resistance of the tube. However, it is to be noted, that in the presence of a current in the winding P_1, and in particular when a condition of balance is attained (which is the state of affairs of primary interest), the flux in the core is reduced to zero and there is no inductive effect in the plate circuit, no magnetic circuit loss and no serious reaction between the primary and secondary circuits involved. The circuital reactions that do exist conspire to preserve the accuracy of the method by their mutual destruction of inductive effects. Few ignorations are therefore necessary and the values of ρ obtained are representative of the tube and not the circuit.

Looking at the thing from another point of view, it is fairly evident that the main current I, flowing thru the winding P_1 induces an emf. in the plate circuit which is of the proper phase to react upon the emf. drop due to the inductance of the winding, P_2. When a perfect balance is obtained, as indicated by silence in the telephones, the neutralization is complete and the plate circuit acts as tho it was devoid of inductance. This may be demonstrated mathematically from the circuital solution, but since space is valuable, reference to the vector diagram of Figure 1 may serve to illustrate the truth of these remarks.

* Received by the Editor, February 18, 1919.

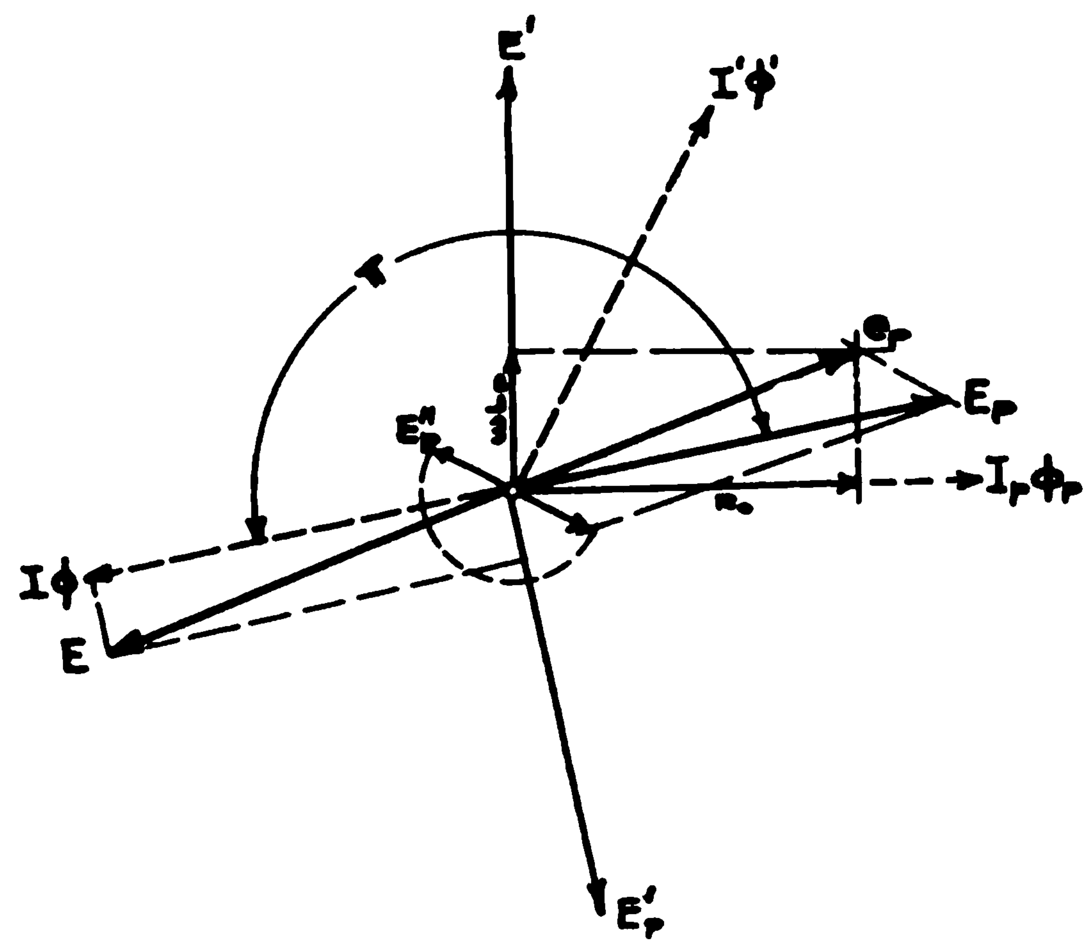

FIGURE 1

In the construction of this diagram no attention has been paid to the proper representation of magnitudes, the phase relations being of paramount interest.

The impressed forces E_p and E are the plate and supply voltages respectively. In addition e_p is the impressed force in the plate circuit on the assumption that there is no reaction from the other primary circuit. When the supply circuit is connected without the emf., e, the flux induces therein the force E' which produces the current, I' and the flux ϕ'. This in turn results in the reactive emf. in the plate circuit, E_p''. The component due to this reaction when combined with e_p gives the final resultant impressed force, E_p. The circuit, as is well known, then acts as if its inductance were reduced and its resistance increased. Due to the connection of the grid across the resistance R in the main circuit, the relation between the main current I and the plate voltage is fixed as shown. The emf. in the main circuit is E; the current produced is I which induces the force E_p' in the plate circuit. Of particular interest here is the fact that this induction is opposed to the normal inductance drop in the plate circuit. This is a crude method of illustrating why the objections mentioned by Dr. van der Bijl tend to disappear when the condition of balance and zero flux is attained. With certain connections, and circuit constants, slight errors due to other causes do exist, as indicated mathematically, but their magnitude is of no practical importance. Furthermore, the

method described makes no pretense at much precision, and is sufficiently accurate for most purposes.

In connection with the other point brought out in Dr. van der Bijl's discussion, I cannot agree with him that the "amplification constant is a function only of the geometry of the tube." It seems to me that the important consideration is the relation between the electronic flow and the forces produced by the electrodes, which is not a simple function of their spatial relation, but depends as has been clearly demonstrated by Richardson, also upon the distribution and congestion of the particles themselves. In constructing the explanation given in the paper to account for the falling off of μ for positive grid potentials I was greatly influenced by this view. Furthermore, I think that a few trial computations will indicate that Dr. van der Bijl's explanation based entirely upon the conductivity of the grid circuit, is quite fallacious. As a matter of fact, this was also the first explanation that occurred to me, suggested probably by the increase in the effect for decreasing plate potentials. The "effective" value of the grid potential is, of course, $\frac{R_1 R_g I}{R_1 + R_g}$ instead of $R_1 I$ as was assumed in my measurements. The slide wire used had a total resistance of about 116 ohms; the resistance of the grid circuit was in no instance lower than 10,000 ohms and in most cases was considerably above this (the exact values are deducible from Figure 11). After correcting for the grid conductance I was surprised to find that the resulting curve was identical with the original one almost within the normal width of the curve itself. I am convinced, therefore, that the effect is negligible and that any explanation based solely upon its consideration is inadequate to account for the pronounced falling off of μ observed. I have but limited faith in the explanation given in the paper to account for this effect and do not wish to contend that this is the correct and final answer to the question. It is frankly crude and explains only the falling off for small plate voltages; the variation for constant plate voltage and variable grid potential still remains unexplained. Any theory for this latter effect must indeed be possessed of remarkable flexibility in order to account for some of the vagaries that have been observed.

In order to give Dr. van der Bijl's theory a fair test I made some very careful measurements of μ with tubes having cylindrical electrodes. This type of constructions gives a relation between the plate current (I_p) and the grid potential (E_g) which

follows most faithfully the relation formulated in Dr. van der Bijl's theory. On account of this agreement with the theory, we should expect the amplification factor to remain constant as the theory predicts. The results are displayed in Figure 2.

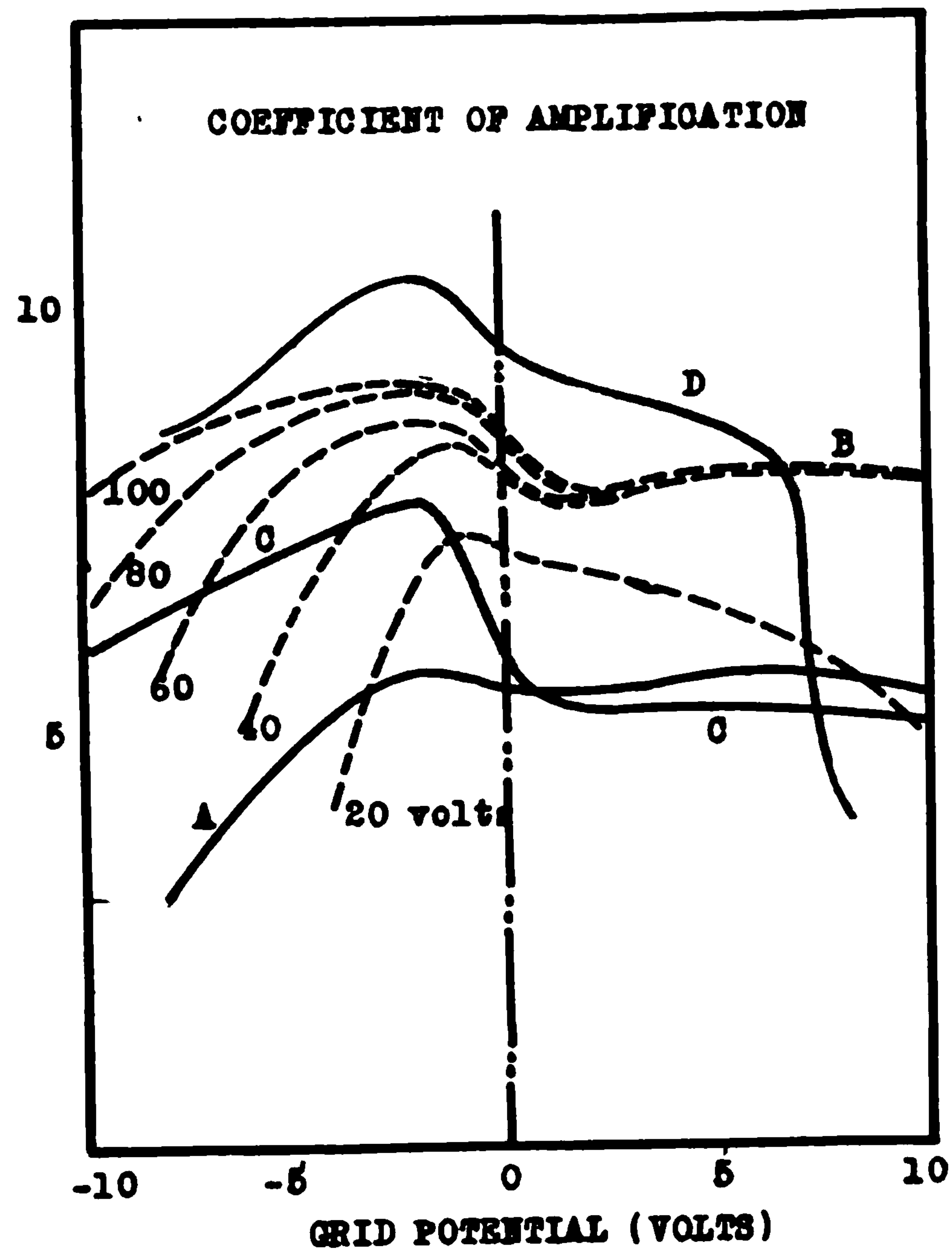

FIGURE 2

The curves "A" and "B" were obtained on tubes with similar plate dimensions and with similar plate-filament spaces. In the case of "A," however, the grid-filament space is greater (by 0.5 millimeter or 0.02 inch) than in the case of tube "B." The falling off of μ for low plate voltages is still noticeable altho not so pronounced as the curves shown in Figure 6, which were obtained with a tube having plane electrodes. The full family of dotted curves is given for tube "B" so that the effect of plate voltage may be indicated. The remaining curves, "C" and "D" were taken with two other tubes of foreign design, tube "D" being of cylindrical form and tube "C" having electrodes of circular form in unilateral arrangement with respect to the filament. All of these curves were taken with the tubes operating

on a plate potential of 80 volts. The very pronounced falling off shown by curve "D" is attributable to an incorrect filament temperature. Raising the filament temperature caused this part of the curve to become horizontal. The filaments of each of the four tubes were made of tungsten and an effort was made to adjust them to the same intrinsic brilliancy. The four curves show a mutual tendency to rise in the region of -2 volts on the grid. After this point is reached, as the grid potential is increased, a sudden fall is noticeable, which is apparently inexplicable and certainly not due to grid conductance since a sensitive galvanometer inserted in the grid circuit shows no evidence of grid current.

I am greatly interested in this matter but feel that too much space has already been consumed in the discussion of an abstract question which is properly a matter of pure science; from an engineering point of view, the variation of μ is not important, particularly in the positive plane. The amplification factor is generally of minor importance in the specification of tube merit, the mutual conductance being a dominant consideration.

Professor Hazeltine's interest in the arrangement shown in Figure 2 is very gratifying. Also his discussion on the influence of the frequency and amplitude of the measuring current are interesting but I have found that in all but very extraordinary cases, these effects are of secondary importance. With gas tubes, in which the phenomenon of impact ionization is not completely negligible, the case is quite different. As first intimated by Vallauri,[1] the hysteresis and viscosity effects then become prominent and the results are very erratic. Fortunately, present day tendencies are directed away from the gas tube, so that this does not become a matter of great concern.

The matter of treating analytically problems relating to vacuum tube circuits seems to resolve itself into a question of individual preference. To one accustomed to the use of straightforward methods of solution by means of differential equations or complex-imaginary algebra (in the case of forced solutions), the conceptions of the coefficient of amplification and internal impedance are of great value. The method which seems to be preferred by Professor Hazeltine involving an "effective" mutual conductance possesses the very important advantage of mathematical simplicity, yet the physical reason-

[1] Vallauri, "L'Eletrotecnica," volume 4, number 18, page 335, 1917.

ing which should parallel any mathematical investigation is much more involved in its nature. In this method, the effect of the reactive emfs. and other extraneous emfs. in the plate circuit require formulation as a correction factor to the normal mutual conductance of the tube. I have always preferred the other method since it seems to be more fundamental and involves the parameters of the tube as fundamental quantities. An example of this mode of dealing with such problems is presented in the treatment of the inductively coupled oscillation circuit given in Addendum 2 to the paper.

Professor Hazeltine has remarked that the two methods of treatment give results which are identical. His execution of the investigation is formally correct, and yet the result he obtains is equivalent to that given in Addendum 2, which is only approximate. The necessity for approximation in my investigation was generated in the solution of the cubic. The extent of the error made is not serious, but it is known to be present. In Professor Hazeltine's treatment no approximations are made so that the lack of rigor becomes apparent. The value of the normal mutual conductance given by Professor Hazeltine is:

$$g_m = \frac{C\,r}{M\left(1 - \frac{1}{\mu} \cdot \frac{M}{L}\right)} \tag{a}$$

which is equivalent to that given in the Addendum:

$$g_m = \frac{C\,r}{M} + \frac{M}{L\,R_p} \tag{b}$$

Proper substitution for μ in (a) transforms it to (b). In other respects Professor Hazeltine's deductions are identical to those which I have made.

The standardization of vacuum tube nomenclature is a matter which is destined to become of increasing importance as this remarkable device is developed. I am inclined to agree with the arguments advanced in Professor Hazeltine's discussion, covering the mutual conductance and internal impedance. The symbol μ for the amplification factor as given originally by Dr. van der Bijl seems to remain satisfactory. The term "internal impedance" is perhaps a misnomer, and should be properly changed to *plate resistance.* The impedance of the tube at radio frequencies is rather complex in form and cannot be represented by a pure resistance as is done at audio frequencies, so that, as Professor Hazeltine remarks, the resistance symbol is quite inappropriate. The symbol R_p is undoubtedly the best suggestion.

I do not feel competent to join in Professor Hazeltine's erudite discussion of the functioning of the tube as a detector in the grid condenser connection. I feel that a derivation of the definition for the merit of the tube is not possible in the absence of an explicit specification of the constants of the associated radio frequency circuit. Even if this information was supplied, the resulting expression involving these circuital coefficients would not be representative of the tube itself, but would represent the operation of the circuit as a whole. The criticism of the definition given in the paper for the detector constant *without* grid condenser is well founded. The result which I gave is erroneous, altho the reasoning based upon Figure 3 is perfectly accurate and leads quite naturally to:

$$D = \frac{1}{2}\frac{d^2 i_p}{d e_g^2} \tag{c}$$

as a definition of the detector constant. The physical structure of the published definition is correct, the error being simply one of quantity. This definition may be checked by assuming the plate current (i_p) to be any function of the grid potential (e_g) as follows:

$$i_p = f(e_g) \tag{d}$$

which is expressible in a power series of the form:

$$i_p = A_o + A_1 e_g + A_2 e_g^2 \cdot \cdot \cdot A_n e_g^n \tag{e}$$

if the coefficients are experimentally evaluated, this may be written in a Maclaurin expansion as follows:

$$i_p = I_o + \frac{d i_p}{d e_g} e_g + \frac{1}{2}\frac{d^2 i_p}{d e_g^2} e_g^2 + \frac{1}{6}\frac{d^3 i_p}{d e_g^3} e_g^3 + \cdot \cdot \cdot \tag{f}$$

The first few terms are of interest only, the indications being that the convergence is complete and rapid in the region in which we are particularly interested. The first term represents the normal plate current at the point of operation; the second term, which is proportional to ρ may be regarded as the amplification. The third term, proportional to the second derivative of the I_p, E_g characteristic apparently represents the rectification and is seen to be identical with (c) above. Succeeding terms may be combined in similar order with the amplification and rectification terms.

In connection with the definition of the detector constant with grid condenser, I do not believe that the phenomenon of detection is quite so simple as indicated by Professor Hazel-

tine's definitions. In constructing the definition which I used in plotting the variation of σ with the grid voltage, my primary object was not to derive something which was quantitatively comparable to the detector constant without grid condenser, but to express something which was of some physical significance and which could be plotted to show the variation of the factor from one part of the characteristic surface to the other. As stated above. an explicit consideration of the quantitative nature of this factor is not possible without analyzing the entire circuit and any expression to possess the desired utility must involve such factors as L, C, R, R_g, and C_g. The magnitude of the detected output is undoubtedly proportional to the extreme excursion of the grid potential in the negative direction regarded as an argument, with the concomitant change in the plate current as the result, this latter effect being proportional to the mutual conductance. The grid potential charge is proportional to the rectification in the grid circuit so that the superposition of these characteristics seems to be of definite physical significance. Professor Hazeltine's remarks evidence a firmer belief in the value of this detecting factor than I have expressed. I have already elaborated my views concerning this so that further discussion is not necessary. If he believes that the grid potential decrease is proportional to the strength of the oscillation and its decrement and is not influenced by the change in detector factor with changing grid potential I do not think that he has given the matter adequate consideration.

Professor Bush's remarks relative to the desirability of being able to specify the performance of vacuum tubes briefly, and without the use of a number of experimental curves, touch upon a very practical matter. It seems to me that the best method of informing the user concerning the capabilities of the device in its various connections as detector, amplifier, modulator, and oscillation generator, would consist in providing each tube with a name plate giving the following information:

	Oscillator	De-tector	Modu-lator	Ampli-fier
Figure of Merit.......	ρ, I_p (*max.*)	D, σ	*	ρ
Plate Resistance......	R_p	..	..	..
Amplification.........	μ	..	..	..
Grid Voltage.........	$\pm$ volts	..	..	..
Plate Voltage.........	+ volts	..	..	..
Filament Current....	Amps.	Amps.	Amps.	Amps.

The factors enumerated are to be measured at the plate and grid potentials corresponding to their maximum values since these are of main interest. This system would not only furnish the figures of merit of the device but would also advise the user of the tube at what part of the characteristic surface to operate. This is obviously desirable and should be helpful in reducing the amount of experimenting necessary to arrange for the proper operating conditions.

* See paper by Dr. J. R. Carson, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, number 2, 1919.

A THEORETICAL STUDY OF THE THREE-ELEMENT VACUUM TUBE*

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The device with the theory of which the present paper is concerned is termed, in default of a generally accepted name, the three-element vacuum tube. Structurally, as is well known, it consists of an evacuated vessel which contains a cathode in the form of an incandescent filament; an anode or plate; and an auxiliary or control electrode which is usually in the form of a grid. The extensive literature which exists concerning this device, and which reflects its scientific and technical importance renders superfluous a description of its structural details or a discussion of the physics of its operation, for a very complete account of which the reader is referred to a recent paper by van der Bijl on the "Thermionic Amplifier."[1]

The purpose of the present paper is two-fold; first, to develop simple formulas which serve as a satisfactory basis on which to construct an elementary theory of the operation of the device in its triple role of amplifier. modulator, and detector, and which indicate the characteristics and factors on which its functioning depends; and secondly, to develop a rigorous mode of dealing with the device by aid of which exact formulas are deducible.

It will be understood that the device dealt with in the following is assumed to be so highly evacuated that the current is transported entirely by electrons emitted from the incandescent cathode, and that ionization of the residual gas molecules plays a quite ignorable part in the mechanism of current conduction.

Secondly it will be assumed that the auxiliary electrode or grid is at all times maintained negative with respect to the

* Received by the Editor, November 12, 1918.

[1] PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, number 2, 1919. This paper also contains a very complete discussion of the theory and operation of the tube as an amplifier.

cathode or filament by means, for example, of suitable "C" battery in the input or grid-filament circuit. When this condition obtains, no current flows in the input circuit and the control is entirely electrostatic.

In the original circuit connections of the audion detector, a condenser was inserted in series with the grid and this arrangement is rather extensively employed to-day. To avoid misunderstanding it should be expressly understood that this arrangement is *not* considered in the present paper which deals with a different mode of detection which does not depend on current rectification in the input circuit.

The physical structure and the fundamental problem involved may be briefly described as follows: The input or grid-filament circuit includes a "C" battery connected with such polarity and of such value as to maintain the grid negative with respect to the filament at all times, and in series therewith a variable emf. corresponding to the impressed effect which it is the function of the device to translate into an amplified, modulated, or detected output as the case may be. The output or plate-filament circuit includes a direct current source of energy termed the "B" battery and in series therewith an external or load impedance z in which is made available the translated output. Corresponding to a variation in the emf. impressed on the input circuit the output circuit current is varied in a manner to be investigated. Our problem, in fact, a solution of which completely predetermines the performance of the device, is to formulate the variation of the output circuit current as a function of the variable input voltage.

With this preliminary understanding we proceed to an analysis of the problem. The basis for this analysis is furnished by the characteristic equation of the tube:

$$I=\phi\left(E_c+\frac{1}{\mu}E_b\right) \qquad (1)$$

where I is the output circuit current; E_c is the instantaneous potential difference existing between the grid and filament; E_B is the instantaneous potential difference between plate and filament; and μ is van der Bijl's "amplification constant" of the device, which depends for its value on the design of the tube and, in particular, on the structure and location of the grid. This characteristic equation, while empiric, holds with great accuracy over the operating range of the device; for a full discussion, see van der Bijl's paper.

Since we are concerned not with the total output current but with its variation in response to a change in the input emf. we write:

$$I=I_o+J$$
$$E_c=E_{co}+e \qquad (2)$$
$$E_b=E_{bo}+v$$

Here obviously I_o may be identified with the steady or normal current in the output circuit corresponding to the steady emfs. E_{co} and E_{Bo}; e is the variation in the input emf. which is to be identified with the variation in the grid-filament potential difference corresponding to the effect to be amplified, modulated or detected; and J is the consequent variation in the output current with which we are concerned. E_{co} and E_{bo} are to be identified with the effective "C" and "B" battery voltages. v is the variation in the plate filament potential difference consequent upon the variation J of the plate-filament or output current. It obviously depends on the value of the load impedance z, and a little consideration will make it clear that v is simply the potential drop, with sign reversed, of the current J thru the impedance z. A clear grasp of this fact is essential to the following treatment of the problem.

Substitution of (2) in (1) and expansion in a power series gives:

$$J=P_1(\mu e+v)+P_2(\mu e+v)^2+P_3(\mu e+v)^3 + P_4(\mu e+v)^4+ \cdots \qquad (3)$$

where $P_1, P_2, \cdots P_n$ are the differential parameters:

$$P_1=\frac{1}{1!}\left(\frac{\partial I}{\partial E_b}\right)_o$$
$$\cdots\cdots \qquad (4)$$
$$P_k=\frac{1}{k!}\left(\frac{\partial^k I}{\partial E_b{}^k}\right)_o$$

The subscript "o" denotes that the derivatives are to be evaluated at the point $E_c=E_{co}$, $E_b=E_{bo}$ of the characteristic.

The necessary conditions that the expansion (3) shall be convergent are

$$E_c<o$$
$$E_c+\frac{1}{\mu}E_b>o \qquad (5)$$
$$I<I_s$$

where I_s is the saturation value of the output current. It will be remarked that conditions (5) simply define the operating range of the tube. The sufficient conditions to insure convergence of the expansion can be determined only when the functional form of ϕ is specified; it appears, however, that in actual tubes ϕ is closely represented by a power of the argument and for this case conditions (5) are sufficient as well as necessary. It may safely be assumed, therefore, as is done in the following discussion, that the expansion (5) is valid over the operating range of the tube.

An investigation of series (3) looking to an explicit formulation of J in terms of e requires a knowledge of v which in turn requires that the load impedance z be specified. The general case where z is unrestricted in form is examined later; for the present we take the simple but illuminating case where z is a pure resistance R. This at once leads to the relation:

$$v = -RJ$$

the substitution of which in (3) gives:

$$J = P_1(\mu e - RJ) + P_2(\mu e - RJ)^2 + \cdots \tag{6}$$

Equation (6) defines J as an implicit function of e, whereas we require its formulation as an explicit function. This is accomplished by inversion of the series (6); the simplest method is to assume an explicit expansion of the form:

$$J = a_1 e + a_2 e^2 + a_3 e^3 + \cdots \tag{7}$$

substitute this series for J in (6), and identify the unknown coefficient by direct equation of like powers of e.

Proceeding in this manner we get the following values for the first three coefficients:

$$\begin{aligned} a_1 &= \frac{\mu P_1}{1 + P_1 R} \\ a_2 &= \frac{\mu^2 P_2}{(1+P_1R)^3} \\ a_3 &= \frac{\mu^3 P_3}{(1+P_1R)^4} - 2R\frac{\mu^3 P_2^2}{(1+P_1R)^5} \end{aligned} \tag{8}$$

These coefficients have a clearer physical significance if we write:

$$\begin{aligned} P_1 &= \left(\frac{\partial I}{\partial E_b}\right)_o = \frac{1}{R_o} \\ P_2 &= \frac{1}{2!}\left(\frac{\partial}{\partial E_b}\frac{1}{R_o}\right)_o = -\frac{1}{2!}\frac{R_o'}{R_o^2} \end{aligned} \tag{9}$$

In accordance with common practice R_o is to be regarded and defined as the internal resistance of the tube or more precisely of the plate-filament circuit.

With this notation:

$$a_1 = \frac{\mu}{R_o + R}$$
$$a_2 = -\frac{1}{2!}\frac{\mu^2 R_o' R_o}{(R_o+R)^3} \tag{10}$$

and

$$J = \frac{\mu e}{R_o+R} - \frac{1}{2!}\frac{\mu^2 R_o' R_o}{(R_o+R)^3} e^2 + a_3 e^3 + \quad . \quad . \quad . \tag{11}$$

The higher order coefficients $a_3, a_4, \quad . \quad . \quad .$ can be evaluated without any trouble; for the present, however, we are concerned primarily with the first two coefficients a knowledge of which is sufficient to construct an elementary theory of the operation of the device. In fact, the higher terms of series (11) represent as may be readily shown, departures from the ideal device in any of its three functions under consideration, and satisfactory operation requires that these higher terms shall be small. As a first approximation we are justified, therefore, in ignoring all terms of the series (11) beyond the second; subsequently, if necessary, the error introduced by their ignoration can be examined. We therefore proceed to a discussion of the operation of the device on the basis of the approximate formula:

$$J = \frac{\mu}{R_o+R} e - \frac{1}{2}\frac{\mu^2 R_o' R_o}{(R_o+R)^3} e^2 \tag{12}$$

In dealing with the problem it is convenient to take the impressed emf. e as

$$A \cos p\, t + B \cos q\, t \tag{13}$$

If we are concerned with amplification the two components of e may be regarded as of the same order of magnitude and comparable frequencies; if modulation is under consideration $B \cos q\, t$ may be regarded as a carrier wave of radio frequency and $A \cos p\, t$ as a signal wave of audio frequency; while as regards detection the form of e given by (13) is appropriate for a study of heterodyne receiving, in which case the two components are to be regarded as both of radio frequency with a frequency difference within the audible range.

Substitution of (13) in (12) and simplification gives:

$$J = \frac{1}{R_o+R}\left\{\mu A \cos p t + \mu B \cos q t\right\} - \frac{1}{2}\frac{\mu^2 R_o' R_o}{(R_o+R)^3}\left\{\begin{array}{l} 2 A B \cos pt \cdot \cos q t \\ +\frac{1}{2} A^2 \cos 2 p t + \frac{1}{2} B^2 \cos 2 q t \\ +\frac{1}{2} A^2 + \frac{1}{2} B^2 \end{array}\right. \tag{14}$$

which may be also written as

$$J = \frac{1}{R_o+R}\left\{\mu A \cos p t + \mu B \cos q t\right\} - \frac{1}{2}\frac{\mu^2 R_o' R_o}{(R_o+R)^3}\left\{\begin{array}{l} A B \cos (q-p) t \\ + A B \cos (q+p) t \\ +\frac{1}{2} A^2 \cos 2pt + \frac{1}{2} B^2 \cos 2 q t \\ +\frac{1}{2} A^2 + \frac{1}{2} B^2 \end{array}\right. \tag{15}$$

Formulas (14) and (15) are fundamental to the following elementary discussion:

Amplification

In amplification the fundamental requirement is that the output shall be a faithful copy of the voltage applied to the input circuit. Consequently, corresponding to an applied voltage as given by (13) the amplified output is to be identified with the term:

$$\frac{1}{R_o+R}\left\{\mu A \cos p t + \mu B \cos q t\right\} \tag{16}$$

of formula (14) while the remainder represents first order distortion.

The outstanding deductions from formulas (16) may be stated as follows:

The effect of impressing a voltage e on the input circuit of a three-element vacuum tube of internal resistance R_o, amplification constant μ and load impedance R, is equivalent, to a first order approximation, to inserting a voltage ($\mu\, e$) in a circuit of resistance R_o+R.

The available amplified voltage across the load resistance is $\frac{\mu R}{R_o+R} e$ and the available output energy is $\frac{R}{(R_o+R)^2}\mu^2 e^2$. The latter is a maximum when $R=R_o$, a result stated by van der Bijl.

As regards the first order distortion, it is proportional to the curvature of the resistance characteristic of the tube. It diminishes with increasing load resistance as stated by van der Bijl; it diminishes also with a decrease of the amplitude of the impressed emf. The curvature of the characteristic, while responsible for the departure of the device from ideal requirements as an amplifier, at the same time makes possible its employment as a modulator and detector.

Modulation

In discussing the phenomena of modulation formulas (14) and (15) are applicable as they stand, but in this case $B \cos q t$ is to be regarded as a carrier wave of radio frequency and $A \cos p t$ as a signal wave of audio frequency. Examination of formula (14) shows that the only terms of the same order of frequency as the carrier wave are:

$$\frac{1}{R_o+R} \mu B \cos q t - \frac{\mu^2 R_o' R_o}{(R_o+R)^3} A B \cos p t \cdot \cos q t \qquad (17)$$

The other terms may be disregarded since, owing to their frequencies, they are suppressed or filtered out by the usual tuning adjustment.

Inspection of (17) shows that the first term is an unmodulated carrier wave of constant amplitude while the second is the modulated output proper; that is, a carrier wave the amplitude of which varies in accordance with the signal wave. As regards the modulated output the obvious deductions from (17) are as follows:

Its amplitude is proportional to the product of three factors

$$(\mu^2 R_o' R_o) \cdot (A B) \cdot \left(\frac{1}{R_o+R}\right)^3$$

The modulated output is therefore proportional to the curvature of the characteristic, R_o', and decreases rapidly as the load resistance is increased. The presence of the factor AB leads to the important practical deduction that the modulated output is independent of the relative amplitudes of the carrier and signal waves provided their product is constant.

The available modulated voltage is proportional to $\frac{R}{(R_o+R)^3}$; this is a maximum when the load resistance is adjusted to make $R = \frac{1}{2} R_o$.

The available modulated energy is proportional to $\frac{R}{(R_o+R)^6}$; this is a maximum when the load resistance is adjusted to make $R=\frac{1}{5}R_o$.

These last two deductions which are valid also for detection, are of considerable practical importance in designing the load impedance for efficient operation. While approximate in that they are based on the neglect of higher terms, it is believed that they are in substantial agreement with the facts. In any case it is to be observed that the presence of higher terms must operate to cause a departure from the ideal modulated wave, so that we are justified in regarding (17) as representing the true modulated wave, the conditions for the maxima of which are substantially as given above.

DETECTION

The phenomena of detection are identical with those of modulation, with the essential distinction that we are concerned with a different order of frequencies in the output; consequently formulas (14) and (15) are adapted as they stand to investigate detection by the heterodyne method. In this case, however, $A \cos p\, t$ and $B \cos q\, t$ are both to be regarded as waves of radio frequency, while the detected output is required to be within the audible range. Consequently discarding the radio frequency terms of (15) the detected wave is made up of:

$$-\frac{1}{2}\frac{\mu^2 R_o' R_o}{(R_o+R)^3}\left\{A B \cos (q-p)\, t+\frac{1}{2}A^2+\frac{1}{2}B^2\right\} \qquad (18)$$

To fix our ideas let $A \cos pt$ be the transmitted wave, and $B \cos qt$ a locally generated wave; then the first term of (18) represents the familiar "beat-note" characteristic of heterodyne reception. The term $\frac{1}{2}B^2$ represents merely a steady value which may conveniently be lumped with the steady current I_o and excluded from explicit consideration. The term $\frac{1}{2}A^2$ represents a change in the normal plate current for the duration of the signal and it alone is cognizable in the absence of the locally generated wave.

Formula (18) furnishes an immediate answer to a question which has been discussed at some length in the pages of this Journal; namely, the theoretical amplification obtainable by

means of a locally generated wave.[2] To answer this question we observe that in the absence of the locally generated wave, the detected effect is proportional to $\frac{1}{2}A^2$, while when a locally generated wave of amplitude B is present the detected current, assuming A as small compared with B, is proportional to $A\,B$. The ratio $\frac{A\,B}{A^2}=\frac{B}{A}$ may be logically regarded as measuring the "heterodyne" amplification, and shows that theoretically it increases without limit as the amplitude of the local wave is increased. Practically, of course, it is limited by the necessity of keeping within the operating range of the tube. The theoretical law is, however, in agreement with the fact that enormous amplification is obtainable by the heterodyne method.

When a modulated wave is to be detected we may identify e of formula (12) with an expression of the form:

$$A\,B\cos q\,t\cdot\cos p\,t+C\cos q\,t$$

where the last term is an unmodulated carrier wave and the first is a wave modulated in accordance with the signal wave $A\cos p\,t$. Substituting in formula (12) and retaining only terms of audio frequency (those comparable with $\frac{p}{2\pi}$ the detected output is:

$$-\frac{1}{2}\frac{\mu^2 R_o' R_o}{(R_o+R)^3}\left\{\begin{array}{l} B\,C\,A\cos p\,t \\ +\frac{1}{4}A^2B^2\cos 2\,p\,t \\ +\frac{1}{4}A^2B^2+\frac{1}{2}C^2 \end{array}\right. \tag{19}$$

Bearing in mind that B and C represent constant amplitudes of the carrier waves, the first term of (19) is directly proportional to the original modulating signal wave; the second term represents a wave of double signal frequency while the last two terms of zero frequency are relatively unimportant and indeed are usually eliminated, as for example, by a transformer. Observe that in the absence of the unmodulated carrier wave of amplitude C, the detected effect is of double frequency; consequently it is obviously desirable to make C large in order to preserve the wave form of the signal wave.

The foregoing discussion is frankly elementary and makes no pretense to being more than a rough approximation to the

[2] See PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS; volume 1, number 3, pages 89, 100; volume 3, number 2, page 185; volume 5, number 1, page 33; volume 5, number 2, page 145; volume 5, number 4, page 247; volume 6, number 5, page 275.—EDITOR.

complicated phenomena. It is believed, however, that it furnishes a good working theory and is in substantial agreement with the phenomena which it represents.

General Solution

We shall now take up the more involved case where the load impedance z is unrestricted in form and is merely any specified function of the frequency. To formulate the solution of this general case we start afresh with the series expansion (3), our problem, as before, being to express J as an explicit function of the impressed voltage e. To make the treatment general, the applied voltage e will be taken as a series of component sinusoids of the form

$$\sum_{k=1}^{k=n} |E_k| \cdot \cos(p_k t + \theta_k)$$

which will be written in the exponential form:

$$e = \frac{1}{2}\sum_{k=n}^{k=n} E_k \, \varepsilon^{j p_k t} + \frac{1}{2}\sum_{k=1}^{k=n} \bar{E}_k \, \varepsilon^{-j p_k t} \tag{20}$$

Here the bar denotes the conjugate imaginary of the corresponding unbarred symbol and the entire expression is, of course, real and equivalent to the cosine summation above.

If the voltage formulated by (20) is applied to a circuit of symbolic or complex impedance z, the resultant current is

$$\frac{1}{2}\sum \frac{E_k}{Z(j p_k)} \varepsilon^{j p_k t} + \frac{1}{2}\sum \frac{\bar{E}_k}{Z(-j p_k)} \varepsilon^{-j p_k t}$$

which will be written as

$$\frac{1}{2}\sum \frac{E_k}{Z_k} \varepsilon^{j p_k t} + \frac{1}{2}\sum \frac{\bar{E}_k}{\bar{Z}_k} \varepsilon^{-j p_k t}.$$

The convenience of this notation will become apparent in the course of the argument. Z is, of course, supposed to be defined as a complex quantity.

Starting with the expansion (3) let us set

$$\begin{aligned} J &= J_1 + J_2 + J_3 + \cdots + J_n + \cdots \\ v &= v_1 + v_2 + v_3 + \cdots + v_n + \cdots \end{aligned} \tag{21}$$

and substitute in (3). The significance of the series formulation of J and v will become apparent in the course of the argument.

At present we observe that the component terms of the

series (21) are unrestricted except by the necessity of satisfying (3) and the relation obtaining between J and v, namely, that v is the potential drop, with sign reversed, of the current J thru the load impedance z.

Now let the components of the J and v series be determined to satisfy the following system of equations:

$$
\begin{aligned}
J_1 &= P_1 (\mu e + v_1) \\
J_2 &= P_1 v_2 + P_2 (\mu e + v_1)^2 \\
J_3 &= P_1 v_3 + 2 P_2 (\mu e + v_1) v_2 + P_3 (\mu e + v_1)^3 \\
J_4 &= P_1 v_4 + 2 P_2 \left[(\mu e + v_1) v_3 + \frac{1}{2} v_2^2 \right] \\
&\quad + 3 P_3 (\mu e + v_1)^2 v_2 \\
&\quad + P_4 (\mu e + v_1)^4 \\
&\cdots\cdots\cdots\cdots
\end{aligned}
\tag{22}
$$

Direct addition by columns of (22), regard being had to the identities of (21), shows that (22) satisfies (3) and is, therefore, a formal solution. The law of formation of the right hand side of each equation will be clear; thus the m^{th} equation includes all terms of order $e^h v^k_{m+1-h-k}$ or all values of k and h from 0 to m. We have further to satisfy the relation obtaining between J and v; this will be accomplished if we identify any v_k with the potential drop corresponding to any component J_k.

Now writing $P_1 = \dfrac{1}{R_o}$ as above, (22) becomes

$$
\begin{aligned}
R_o J_1 - v_1 &= \mu e \\
R_o J_2 - v_2 &= R_o P_2 (\mu e + v_1)^2 \\
R_o J_3 - v_3 &= 2 R_o P_2 (\mu e + v_1) v_2 + R_o P_3 (\mu e + v_1)^3 \\
&\cdots\cdots\cdots\cdots
\end{aligned}
\tag{23}
$$

Bearing in mind the assumed physical significance of the v series, it follows at once that any equation say k^{th} is simply that of the current J_k in a circuit of impedance $R_o + z$ in response to a voltage given by the right hand side of the equation. The recognition of this fact furnishes the key to the complete formal solution for J.

To carry this out substitute e as given by (20), in the first of (23); then it follows at once that

$$J_1 = \frac{1}{2}\mu \sum \frac{E_k}{R_o + z_k} \varepsilon^{j p_k t} + \frac{1}{2}\mu \sum \frac{\bar{z}_k}{R_o + \bar{z}_k} \varepsilon^{-j p_k t}$$

If we write

$$R_o+z_k=Z_k$$

$$\frac{z_k}{R_o+z_k}=\rho_k$$

then:

$$J_1=\frac{1}{2}\sum\frac{\mu E_k}{Z_k}\varepsilon^{jp_kt}+\frac{1}{2}\sum\frac{\mu \bar{E}_k}{\bar{Z}_k}\varepsilon^{-jp_kt}$$

and (24)

$$v_1=-\frac{1}{2}\sum\mu\rho_k E_k\varepsilon^{jp_k t}-\frac{1}{2}\sum\mu\bar{\rho}_k\bar{E}_k\varepsilon^{-jp_k t}$$

Equations (24) state that to a first order approximation the effect of impressing a voltage e on the input circuit of a tube of amplification constant μ, internal resistance R_o and load impedance z, is equivalent to inserting a voltage μe in a circuit of impedance $R_o+z=Z$. This is the generalized form of the law already stated when z is a pure resistance R and shows that this restriction is unnecessary. J is, of course, to be identified with the amplified output current of the ideal amplifier.

By virtue of (24) and (20), the right hand side of the second equation of the system (23) is known; it is after easy simplification:

$$\frac{\mu^2}{4}R_o P_2\sum^h\sum^k\begin{cases}(1-\rho_h)(1-\rho_k)E_hE_k\varepsilon^{j(p_h+p_k)t}\\+(1-\rho_h)(1-\bar{\rho}_k)E_h\bar{E}_k\varepsilon^{j(p_h-p_k)t}\\+(1-\bar{\rho}_h)(1-\bar{\rho}_k)\bar{E}_h\bar{E}_k\varepsilon^{-j(p_h+p_k)t}\\+(1-\bar{\rho}_h)(1-\rho_k)\bar{E}_hE_k\varepsilon^{-j(p_h-p_k)t}\end{cases}\tag{25}$$

The last two terms of the double summation are the conjugate imaginaries of the first two, respectively. This expression may be regarded as the applied emf. which generates the current J_2; consequently J_2 and v_2 are given by

$$J_2=\frac{\mu^2R_oP_2}{4}\sum_{h=1}^{h=n}\sum_{k=1}^{k=n}\begin{cases}\dfrac{(1-\rho_h)(1-\rho_k)}{Z_{h+k}}E_hE_k\varepsilon^{j(p_h+p_k)t}\\+\dfrac{(1-\rho_h)(1-\bar{\rho}_k)}{Z_{h-k}}E_h\bar{E}_k\varepsilon^{j(p_h-p_k)t}\\+\dfrac{(1-\bar{\rho}_h)(1-\bar{\rho}_k)}{\bar{Z}_{k+k}}\bar{E}_h\bar{E}_k\varepsilon^{-j(p_h+p_k)t}\\+\dfrac{(1-\bar{\rho}_h)(1-\rho_k)}{\bar{Z}_{h-k}}\bar{E}_hE_k\varepsilon^{-j(p_h-p_k)t}\end{cases}$$

(26)

$$v_2 = -\frac{\mu^2 R_o P_2}{4} \sum_{h=1}^{h=n} \sum_{k=1}^{k=n} \begin{cases} \rho_{h+k}(1-\rho_h)(1-\rho_k)\, E_h E_k\, \varepsilon^{j(p_h+p_k)t} \\ +\rho_{h-k}(1-\rho_h)(1-\bar{\rho}_k)\, E_h \bar{E}_k\, \varepsilon^{j(p_h-p_k)t} \\ +\bar{\rho}_{h+k}(1-\bar{\rho}_h)(1-\bar{\rho}_k)\bar{E}_h\bar{E}_k\, \varepsilon^{-j(p_h+p_k)t} \\ +\bar{\rho}_{h-k}(1-\bar{\rho}_h)(1-\rho_k)\bar{E}_h E_k\, \varepsilon^{-j(p_h-p_k)t} \end{cases} \tag{27}$$

In these summations Z_{h+k} denotes the complex impedance of the circuit R_o+z to the frequency $\frac{p_h+p_k}{2\pi}$; Z_{h-k} denotes that to frequency $\frac{p_h-p_k}{2\pi}$; while in accordance with the notation adopted

$$\rho_{h+k} = \frac{z\,(j\,(p_h+p_k)\,)}{R_o+z\,(j\,(p_h+p_k)\,)} = \frac{z_{h+k}}{R_o+z_{h+k}}$$

With v_2 determined by (27) the right hand side of the third member of the system (23) is known and consequently the third order components J_3 and v_3 can be written down. Owing, however, to the complexity of the resulting expressions a more compact notation is required, and the following symbolic notation commends itself for this purpose. If we let a negative subscript attached to any symbol denote the conjugate imaginary of the same symbol with positive subscript and extend the summations over negative as well as positive values, we write

$$e = \frac{1}{2} \sum_{k=-n}^{k=n} E_k\, \varepsilon^{j p_k t}$$

which is equivalent to (20). In this same notation J_1 and v_1 are given by

$$J_1 = \frac{\mu}{2} \sum_{k=-n}^{k=n} \frac{E_k}{Z_k} \varepsilon^{j p_k t}$$

$$v_1 = \frac{-\mu}{2} \sum_{k=-n}^{k=n} \rho_k\, E_k\, \varepsilon^{j p_k t}$$

while J_2 and v_2 become:

$$J_2 = \frac{\mu^2 R_o P_2}{4} \sum_{h=-n}^{h=n} \sum_{k=-n}^{k=n} \frac{(1-\rho_h)(1-\rho_k)}{Z_{h+k}} E_h\, E_k\, \varepsilon^{j(p_h+p_k)t}$$

$$v_2 = -\frac{\mu^2 R_o P_2}{4} \sum_{h=-n}^{h=n} \sum_{k=-n}^{k=-n} \rho_{h+k}(1-\rho_h)$$
$$(1-\rho_k)\, E_h\, E_k\, \varepsilon^{j(p_h+p_k)t}$$

These last two equations are equivalent to (26) and (27) respectively.

The right hand side of the third member of (23) becomes:

$$\frac{\mu^3 R_o}{8} \sum_{g=-n}^{g=n} \sum_{h=-n}^{h=n} \sum_{k=-n}^{k=n} (P_3 - 2R_o P_2^2 \rho_{h+k})(1-\rho_g)(1-\rho_h)$$
$$(1-\rho_k)\, E_g\, E_h\, E_k\, \varepsilon^{j(p_g+p_h+p_k)}$$

by aid of which J_3 is given by:

$$\frac{\mu^3 R_o}{8} \sum_{g=-n}^{g=n} \sum_{h=-n}^{h=n} \sum_{k=-n}^{k=n} \frac{P_3 - 2R_o P_2^2 \rho_{h+k}}{Z_{g+h+k}} (1-\rho_g)(1-\rho_h) \cdot$$
$$(1-\rho_k)\,.\,(E_g E_h E_k\, \varepsilon^{j(p_g+p_h+p_k)t}$$

It is to be regretted that the complexity of the formulas require such an abbreviated notation, but this is inherent in the nature of problem and I know of no other system of symbolic notation by which the results can be written down in compact form. As a check on this solution we can put the load impedance z equal to a pure resistance R, in which case the general solution degenerates into a term by term identity with the series solution derived in the first part of this paper. The complexity of the general solution is of course due to the presence of harmonic frequencies and the fact that the load impedance is a frequency function instead of being a constant.

The solution, it should be observed, while formally correct, is incomplete without an investigation of the convergence of the J and v series, a rigorous discussion of which is beyond the scope of this paper. It can be shown, however, that the formal solution is a Fourier series, the coefficient of each term of which is an infinite convergent series within the range of convergence of the expansion (3). The range of validity of the formal general solution is, therefore, defined by the region of convergence of the original expansion from which it was derived.

September 9, 1918.

SUMMARY: Starting with a functional relation between the plate current and the grid and plate voltages, the plate current variation is deduced to a sufficient degree of approximation in terms of constants of the tube and the grid voltage variation (for pure resistance in the output circuit).

The requirements of amplification, modulation, and detection are then deduced from the equations obtained. Heterodyne amplification is discussed, and it is held that within the working range of the tube this amplification is the ratio of the amplitudes of the locally and remotely generated waves provided the latter is small as compared with the former.

The general problem of output current variation with an impedance in the output circuit is then considered, and the production of harmonic frequencies noted.

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RECEPTION THRU STATIC AND INTERFERENCE*

BY

ROY A. WEAGANT

(CHIEF ENGINEER, MARCONI WIRELESS TELEGRAPH COMPANY OF AMERICA, NEW YORK)

Since the birth of radio telegraphy, serious difficulty in reception has existed due to natural electrical disturbances. These disturbances produce in the receiving telephones crackling noises which drown out the signal and are commonly called static, atmospherics, or strays. In what follows the word "static" will in general be used in referring to these disturbances, whatever their nature or origin.

As the distance over which radio telegraphy was worked increased, and it became necessary to use increasingly longer wave lengths, it was found that the troubles from static continually increased and in the case of the most important of long distance circuits, namely those between Europe and the United States, caused such great interruptions to the service that the continuity of communication compared very poorly with that of cable working. It was found that static disturbances were most severe in summer and less troublesome in winter, also that they displayed a daily variation in intensity, being at a minimum between sunrise and noon, and increasing very rapidly to a maximum about sunset, from then on remaining practically constant until shortly before sunrise when the intensity fell off very sharply to a minimum again.

Accumulated experience shows that these disturbances are more severe in locations near or in the tropics than in those of the temperate zone or frigid zone, and also that at any given location they vary from day to day somewhat in accordance with the variations in temperature, being greater on warm days and less on cool days as a rule, altho not invariably so.

A great deal of study has been made in attempts to determine the nature and origin of these disturbances and innumerable attempts to secure methods of reducing their effects at the re-

* Received by the Editor, March 4, 1919. Presented before THE INSTITUTE OF RADIO ENGINEERS, New York, March 5, 1919.

ceiving station, but so far as the writer is aware, no success of a major order was obtained with any of these methods prior to the work which is about to be described. One of the most common of these previous arrangements known to the writer made use of a transmitter of undamped waves and beat reception. How far short of meeting the situation in trans-Atlantic working this method fell, may be judged from the fact that from June to October good reception from such continuous wave stations as Carnarvon, Wales, and Nauen, Germany, was usually possible only between sunrise and noon, while during the rest of the day it varied from very poor to totally impossible. An idea of the magnitude of the problem to be met can be gathered from the fact that during these summer months the energy collected by a receiving aerial from static is often many thousands of times as great as that of the normal signal from the above-mentioned stations.

It is a well recognized fact that static disturbances are of different sorts which are apparently due to a variety of causes, and of these different varieties those due to local lightning and snowstorms will be dismissed for the present with the statement that they occur so infrequently as to be of negligible consequence. There then remain three other major types which have been generally recognized and which Eccles has classified under the names of "grinders," "clicks," and "hissing." The last of these types, due generally to an actual discharge from antenna to earth, produces very little disturbance and is not present when antennas are used which have no earth connection. Of the two remaining types, namely, the grinders and the clicks, it is found that the former constitute the major source of difficulty in the reception of trans-Atlantic signals, the intensity of which is that of Nauen or Carnarvon, and when the receiving station is located in the United States. It is this form of static which rises to overwhelming intensity in the summer months and which has hitherto produced such serious interruption in trans-Atlantic radio communication. It should be noted, however, that both types of static are generally present, but that as the grinders increase in intensity, in general the clicks diminish. As will develop in the course of this paper, these two types of static are, apparently, of totally different nature and origin.

To make clear the various steps in the developments which are to be described, reference to certain fundamental facts, which are a matter of common experience in radio reception, is neces-

sary, and, briefly, to various methods of overcoming static troubles which have been tried. In this latter respect it is not to be understood that an exhaustive statement of all the various methods of solution is presented nor an accurate comparison of their relative values, but merely such reference as is necessary in order to trace the steps of the writer's work. It is also to be understood in what follows that the major portion of the work which is here referred to has been with signals from Europe of wave lengths varying between 5,000 and 15,000 meters. Some work has also been done with shorter wave lengths and the results secured were in substantial agreement with those obtained in the range above mentioned, but this work has not been of an exhaustive nature.

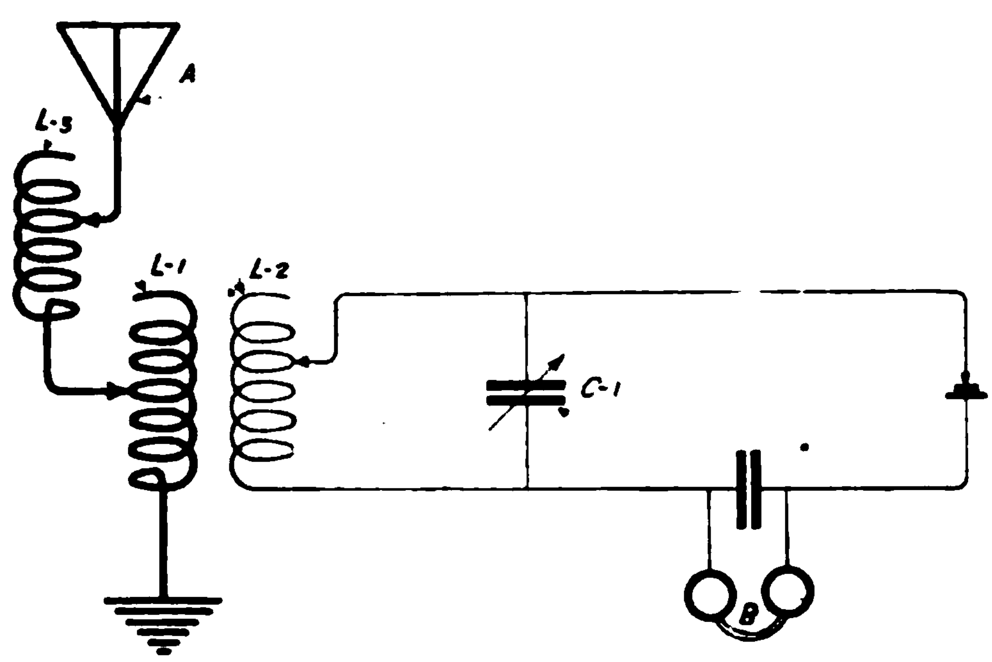

FIGURE 1

Referring now to Figure 1, there is outlined a simple form of common receiving system, the elements of which need no description. When such a system is tuned and adjusted to give best response to the incoming signal it is found that the disturbances from static are also invariably a maximum, regardless of the frequency to which the system is adjusted. A study of the behavior of such a system when acted upon by static very clearly brings out the fact that the disturbing currents which flow therein have a period and damping which is determined by the circuit itself; a fact which shows that the disturbance is in the nature of a shock, the system, when so shocked, vibrating in a way which is analogous to that of a tuning fork struck by a hammer.

It is curious to note the number of experimenters who, while apparently recognizing this principle, immediately attempt

to secure relief from static disturbances by detuning methods which result simply in the reduction of both signal and static currents in substantially equal proportion, and consequently with no appreciable improvement. This result is due to the fact that while detuning the aerial circuit does not reduce the intensity of the static in the antenna circuit, it does change the frequency of the currents due to it; and the loss in transfer of energy to the secondary circuit, since the latter is tuned to the frequency of the incoming signal and therefore a different frequency from the detuned antenna, is of exactly the same order as the loss in intensity which the signal currents experienced when the antenna circuit was detuned. Another simple expedient which has been resorted to, has been the employment of loose couplings between the antenna circuits and the secondary circuits, and this method does give some help when the difference in damping between the signal currents and static currents is marked. Attempts to make this difference as large as possible have been made, involving the introduction of resistance into the antenna and secondary circuits, but this always results in the reduction of both signal and static currents by a substantially proportional amount, with a resulting negligible order of improvement. A large number of arrangements with which it was hoped to secure differentiation, and depending on this principle of difference in damping of the two currents involved, have been tried but, so far as is known to the writer, without important results.

Another fundamentally incorrect method of attack is that of differentially combining two circuits, of which the Fessenden interference preventer circuits shown in Figure 2 are typical.

The antenna circuit here shown is split into two branches, each coupled to a common secondary and detector circuit, or these individual branches may be connected to two different antennas. One of the branch circuits was supposed to be detuned slightly with respect to the incoming signal, materially reducing the signal current in that branch, but not appreciably affecting the static current, which was assumed to be a *forced* oscillation and which would not therefore have either its frequency or intensity affected by an amount of detuning which would greatly affect the signal. The remaining static currents would then, supposedly thro the common coupled circuit connected in opposition, cancel the static due to the other branch, leaving a signal current equal to the difference between that existing in the two branches. Several other methods of adjust-

ment were proposed, among which was that of adjusting one branch to a period slightly below the incoming signal, and the other to a period slightly above it. There are many fallacies

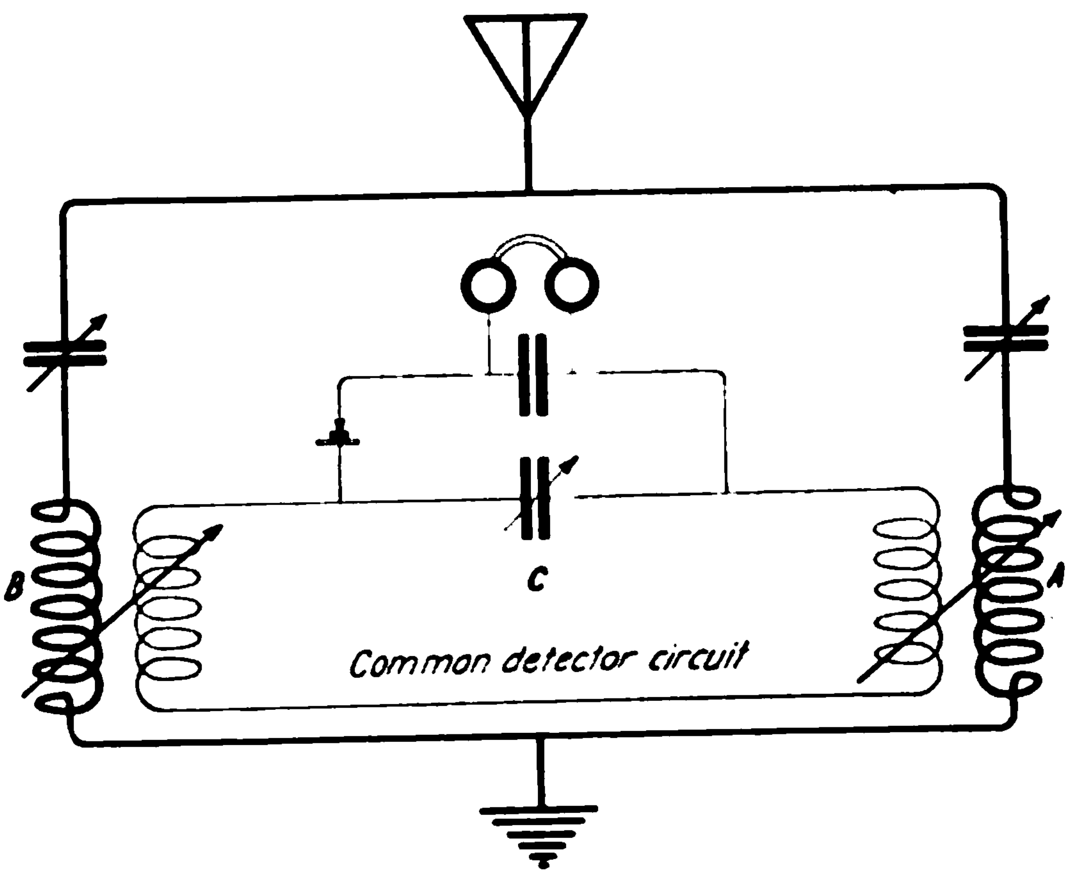

FIGURE 2

in this proposal, but it is sufficient for present purposes to point out the facts stated in connection with Figure 1, namely, that the detuning of one branch circuit affects the intensity of both the signal and static currents in the secondary circuit in the same ratio, and the additional fact that if one circuit is tuned to a period differing from that of the other, the frequency of the static in the first named circuit is different from that in the other circuit, and two alternating currents of different frequencies obviously cannot neutralize each other, but on the contrary, in order that such neutralization may be accomplished it is necessary that the emfs. which are to equalize each other must be of the same frequency, the same wave form, and of opposite phase. Also when these emfs. are due to the flow of damped oscillating currents, these currents must have the same damping factor. If this requirement is complied with in the arrangement of Figure 2, the static currents will cancel out but so also will the signal currents. Many variations of the arrangement of Figure 2 have been tried, including some in which the differentiation is attempted in the audio frequency instead of the radio frequency circuits, but if any of them have seemed to work, the result secured has been entirely due to the looseness of coupling involved.

is to be noted, however, that under some circumstances such an aerial may equally well be acting as a loop; such an aerial is shown in Figure 5 lying on the surface of the ground and it is evident that by virtue of its capacity to the true conducting earth, a return path between its ends exists and therefore that it is a form of loop; which method of consideration will account for many of the observed facts, such as its directivity, in a satisfactory way. It will also account for one observed fact which

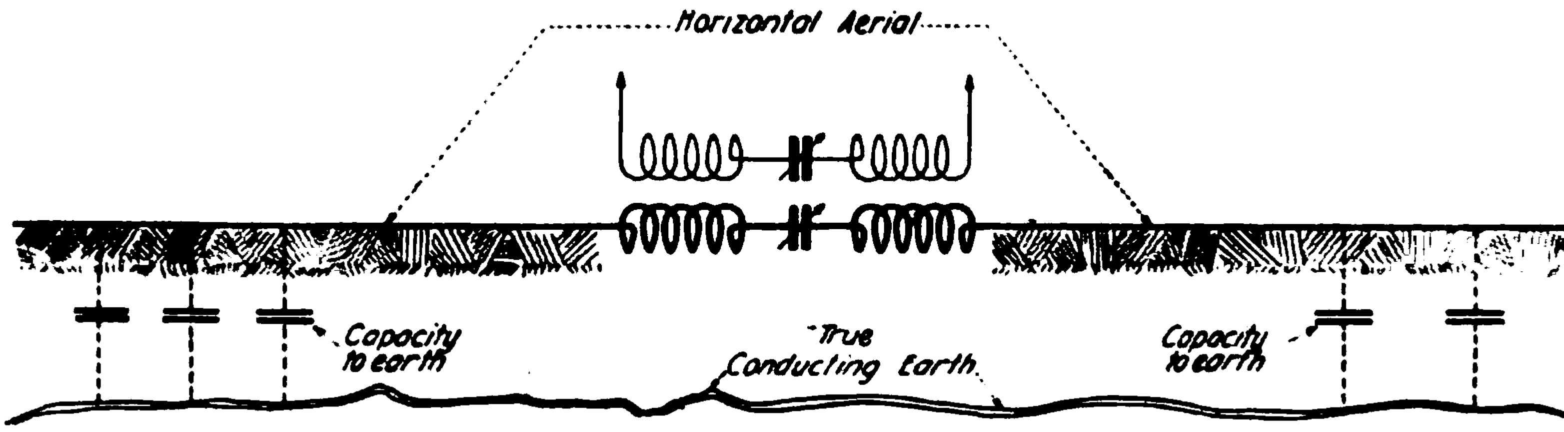

FIGURE 5

the usual methods of explanation do not account for, namely, that when an aerial of this type is laid on the ground, or buried underneath it, its effectiveness as an antenna does not increase indefinitely with length but rapidly reaches an optimum value dependent on the circumstances obtaining. This can readily be accounted for under the present hypothesis by the fact that as the length increases its capacity to earth increases and at some point becomes sufficient to close the loop.

As this capacity increases, however, the currents originating in the increased length have various paths in which to flow, one of which includes the receiving apparatus, but others which are thru the capacity to earth between the conductor and the receiving apparatus, and the larger this gets the greater is the proportion of the currents originating in the ends of this antenna, which are diverted and do not flow thru the receiving apparatus. This method of considering such an antenna is further supported by the fact that the greater the capacity per unit of length which exists between the conductor and the true underlying earth, the shorter is the maximum length which can be used to advantage. This capacity is a maximum of course when the antenna is actually buried in the ground or under water, becoming less when the wire is run on the surface of the earth and still less when the wire is suspended at some height above the earth, tests having shown that wires suspended some 10 feet

(3 m.) above ground can be used up to some six miles (9.6 km.) in length, the signal increasing with length; that a length about one-half of this is effective when the wire is laid on the ground and of approximately 2,500 feet (760 m.) when the wire is placed under brackish water.

I have also found that as the distance of such an antenna above ground is increased, its action becomes more nearly that of an ordinary antenna, and that therefore on account of its position relative to the incoming signal, it becomes less effective in collecting this signal energy.

While the two forms of antennas just referred to result in a distinct and important advance over other types of antennas, the improvement in results secured there from falls very greatly below that which is necessary to meet the conditions of continuous trans-Atlantic reception.

Another method of attack is the screen arrangement suggested by Dieckmann and de Groot which has no basis, so far as the writer can see, for differentiating between static and signal, but must, if it has any effect at all, operate on both alike. Furthermore, in attempting to investigate screening arrangements of this sort, it has been found that the problem of screening out an electro-magnetic wave of any sort, either signal or static, is not solved by the methods mentioned by them.

One of the most important investigations of static effects was that carried out by Mr. C. H. Taylor, of the Marconi Company, in which the Bellini-Tosi direction finder arrangements were used in an attempt to find out in what, if any, horizontal direction static disturbances were propagated. Altho this work showed that at times there was some definite evidence of direction of propagation, it did not warrant the hope that a successful method of separation could be based thereon. The writer's observations made at this time, and with the same installation used by Mr. Taylor, and with a similar arrangement erected at the Marconi Company's New Brunswick station, showed that so far as the dominant type of static—namely the grinders—was concerned, no direction whatever could be found, but on the contrary there appeared to be an equality of disturbances from all points of the compass. A further check on this result was made at this time by rotating a loop, shown in Figure 6, about a vertical axis; this also showed equality of average disturbances, regardless of the direction of the plane of the loop and led to the conclusion that if static disturbances of the grinders type were being propagated horizontally, they

must be moving in all possible directions; that is to say, one stroke might arrive from the north, the next one from the east, a third from the west, and so on, these occurring at random in such rapid succession as to give no opportunity to deter-

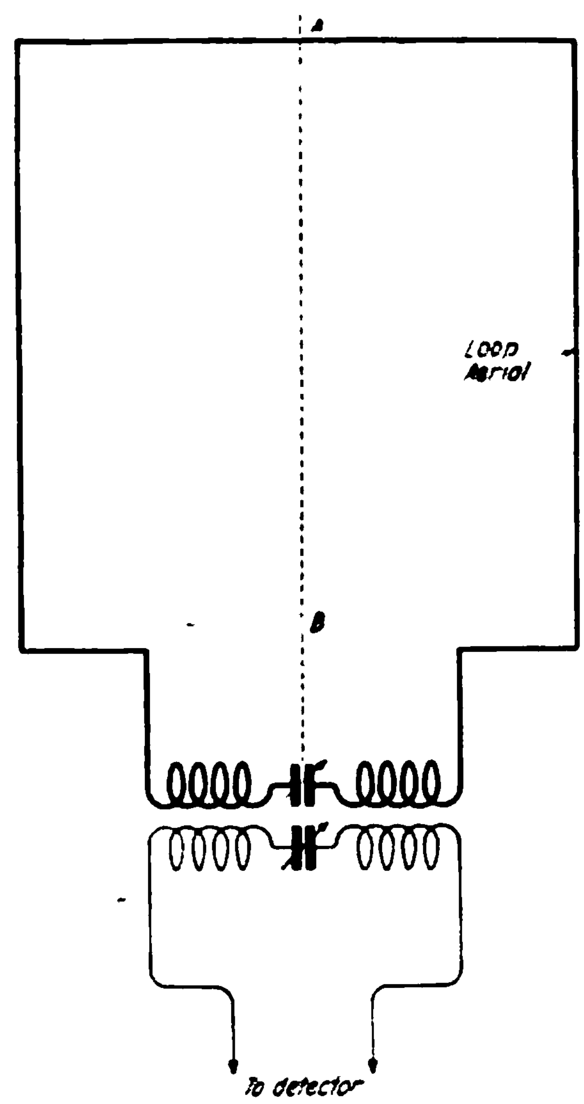

FIGURE 6

mine their direction. There appeared to the writer, however, another possible explanation of this result, which is that these disturbances instead of moving horizontally, might be moving in a vertical direction, the source being under foot or overhead. This latter possibility was of exceptional interest since, if it were correct, the direction of propagation of static waves, assuming of course that they were waves, would be at right angles to the direction of propagation of signals, and such a difference might conceivably be used to separate the two.

Steps were then taken to determine which of the two possible explanations given above was correct, and the investigation seemed to establish clearly and definitely that static of the grinders type produces effects similar to those which would be produced by electro-magnetic waves originating overhead or under foot and propagated in a direction perpendicular to the earth's surface at the point of observation. This investigation also established that static currents produced

in loops, the planes of which are perpendicular, cannot be combined to neutralize each other, which result can be explained by assuming that the electro-magnetic waves responsible for static currents are heterogeneously polarized; that is, the axes of the oscillators producing them assumed all possible angles in space. To sum up then, *these results showed that static disturbances of the grinders type behaved as tho due to heterogeneously polarized, electro-magnetic highly damped waves propagated in a direction perpendicular to the earth's surface.*

The apparatus and method used in this investigation resulted in a perfectly practical receiving system which, while retaining useful amounts of signal currents, enormously reduced the currents due to static of the dominant type. The methods and apparatus used in carrying out these tests were as follows:

Two single turn loop antennas were erected 400 feet (122 m.) high each with a base line of 1,000 feet (305 m.) and their centers approximately 5,000 feet (1,520 m.) apart. These loops were in the same plane and the line connecting them was in a direction toward the Carnarvon station of the English Marconi Company. This arrangement is shown schematically in Figure 7. Leads

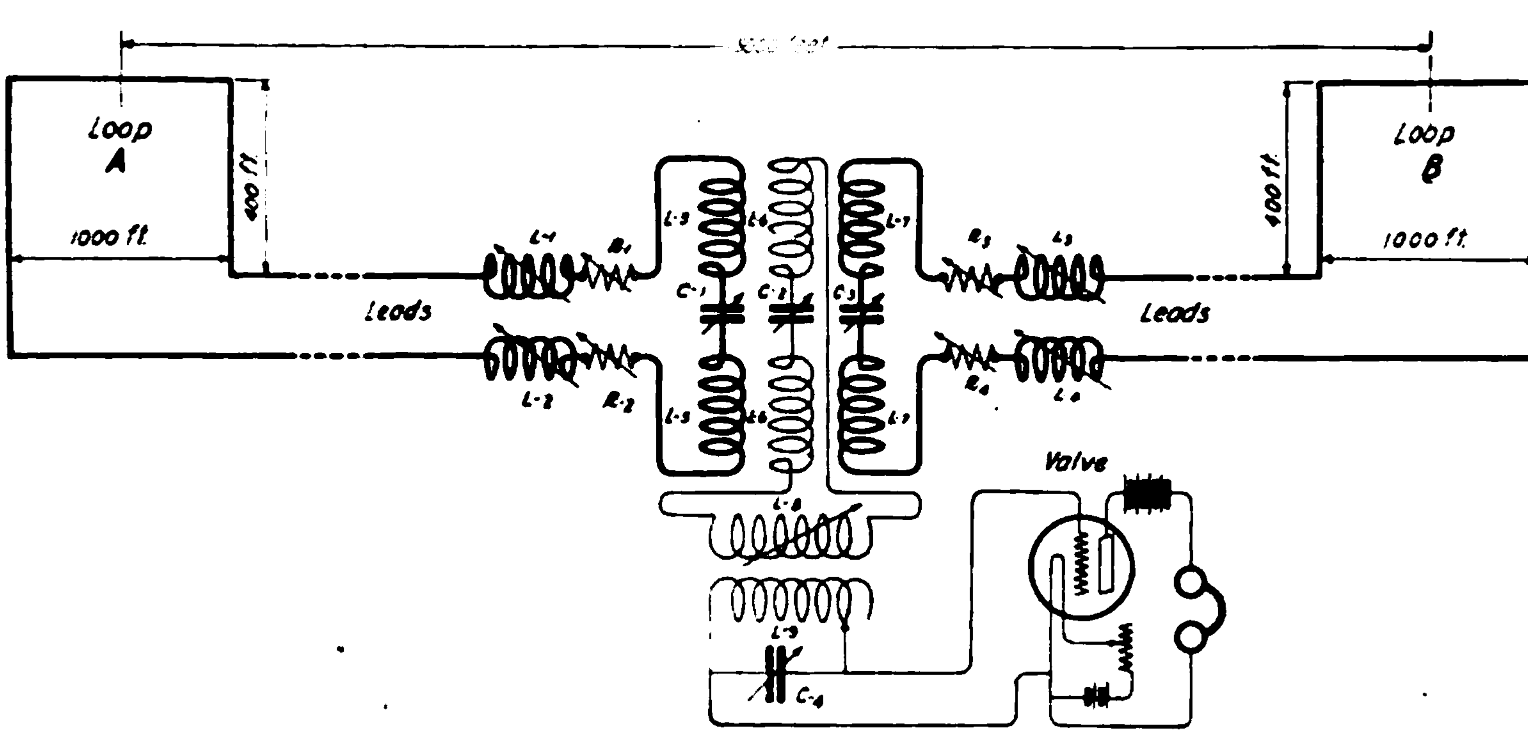

FIGURE 7

were brought from the loops to a receiving station midway between them. These leads, which were six feet (1.82 m.) apart, and in the same horizontal plane, were supported by poles about ten feet (3.05 m.) high. The diagram of connections is shown in Figure 7.

Connection from the leads were made thru inductances L_1, L_2, L_3, and L_4 symmetrically arranged relative to the coils

L_5 and L_7, which were arranged perpendicularly to each other, as shown in Figure 8. The winding of each fixed coil L_5 and L_7 was divided into two equal parts, and condensers C_2 C_3 inserted between the halves. Associated with the two fixed

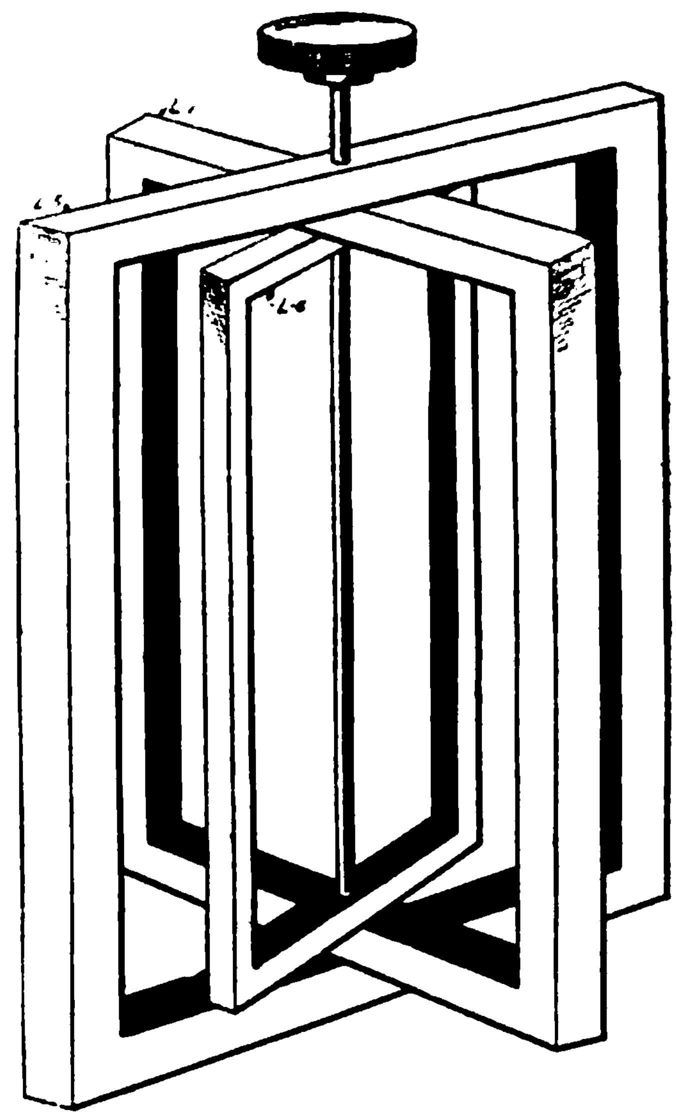

FIGURE 8

coils was a third coil L_6 capable of rotation on a vertical axis, the three coils constituting the well-known Bellini-Tosi goniometer. In the circuit containing L_6 were condensers C_2 and coupling coil L_8 which was associated with a receiver of conventional type with valve detector. The theory of the tests made with this arrangement is as follows:

Assuming that static waves were traveling perpendicularly to the earth's surface, then the electromotive forces generated in the two loops would be equal in intensity and of the same direction at any instant, and therefore if the circuits were properly tuned, the resulting currents in the system would be in phase. The emfs. generated by the signal would, on the other hand, be out of phase by an amount depending on their distance apart, and a maximum if this distance were one-half the length of the wave received, since the signal wave would arrive at the antenna nearest the transmitting station before it would arrive

at the antenna farthest away from the transmitting station. In other words then, the static waves would arrive at the two antennas at the same time, while the signal waves would arrive at the two antennas at different times. It therefore follows that if, at the receiving station, connections and adjustments were so made that the emfs. generated in the rotating coil L_6 by static disturbances were equal and opposite, the emfs. generated by the signal currents would not be equal and opposite, but would combine, giving a resultant depending on the separation of the loops. If this separation were one-half wave length then the emfs. generated in coil L_6 by the signal currents from each loop would be in phase and would therefore be equal to the arithmetical sum of these two emfs. If the loop separation were equal to one-quarter of a wave length, then the emfs. acting on the coupling coil would be 90 degrees apart and the resultant would be equal to 1.4 times that of the individual emfs.; that is, they would continue in quadrature. If, on the other hand, the hypothesis that static of the grinders type arrives from all possible azimuthal angles, and in a horizontal direction, were correct, then the static currents arriving at the receiving station from the two antennas would be out of phase an amount depending on the separation of the antennas and the azimuthal angle which the direction of their propagation made with the base line of the system.

If the apparatus in the receiving station were assumed to be adjusted in such a way that the signal currents were combined vectorially and in accordance with the aerial separation, then the static currents would be similarly combined; and the curve of reception of the system so adjusted for static impulses equally distributed in all azimuthal angles would be that of Figure 9, which is nearly the same as that of a single loop antenna, and therefore the whole system would show nearly the same signal-to-static ratio as the single loop. If, on the other hand, adjustments were so made that the phases of the currents from one loop were shifted a suitable amount with respect to the phases of the current in the other loop, then the curve of reception would change from that of Figure 9 to that of Figure 10, which indicates that reception through one-half of the azimuthal angles has been moderately reduced, and which would therefore give rise to an improvement in the signal-to-static ratio of the whole system as compared with the single loop of the order of the decrease of the area included by the curve of Figure 10. It will thus be seen that under the three sets of conditions specified

and under the two hypothesis considered, there were three possible results, namely, a very large improvement in the signal-to-static ratio under the first hypothesis, a small improvement if the second hypothesis were correct, and the first method of adjustment followed, and a moderate order of improvement under the second hypothesis and the second method of adjustment.

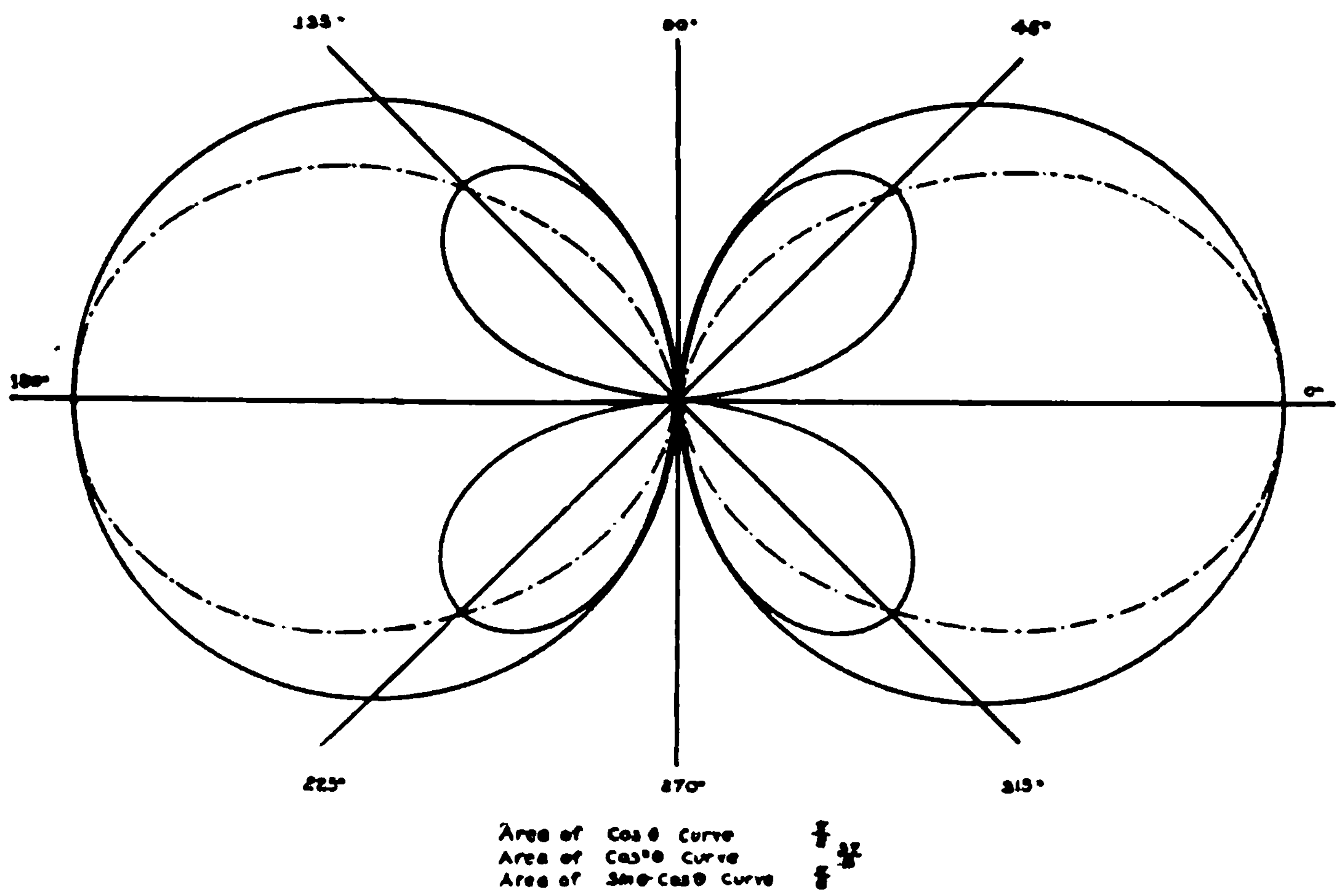

FIGURE 9

Referring now to the arrangement actually used, the spacing between the loops was slightly over one-quarter wave length, for a wave length of 6,000 meters, which was that used by Nauen during some of the tests. Signals were also received from Nauen

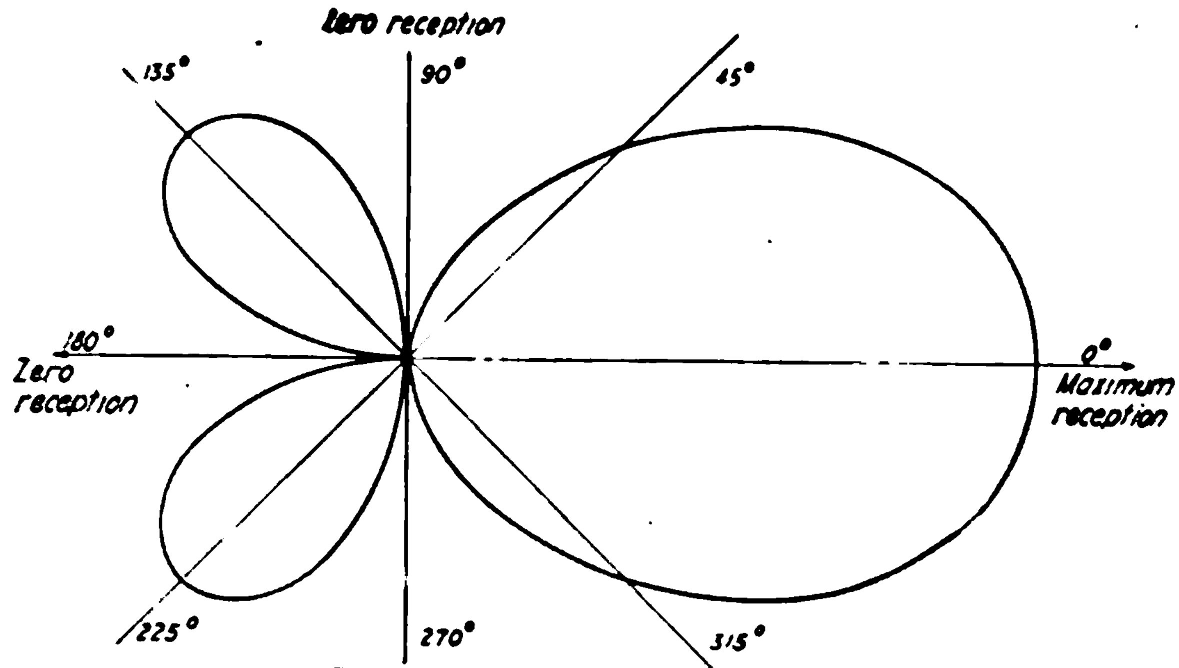

FIGURE 10

at 12,000 meters, Clifden 5,600 meters, Carnarvon 14,000 meters, Eilvese 9,600 meters, and Glace Bay 7,600 meters. In all cases it was found that when the adjustment of the circuits was so made that static disturbances of the grinders type were cancelled or reduced to a minimum, the signal received from the two loops combined, as might be expected from the spacing between them, and the wave length of the incoming signal. In the case of the 6,000-meter signal from Nauen, the resultant signal was approximately forty per cent greater than that due to either aerial alone, while in the case of Carnarvon, with a 14,000 meter wave length, for which the spacing was equal to only one-nineth of a wave length, the resultant signal was materially less than that due to either loop alone.

Since the order of improvement of the signal-to-static ratio of the system as a whole was very great as compared to the single loop, and consequently a given signal was readable thru static disturbances of a very much greater order than was possible with the single loop, it was concluded that the hypothesis of an apparent vertical propagation of static waves (or an electric action of equivalent effect) more nearly expressed the true facts than did that which assumed a uniform azimuthal distribution of their horizontal direction of propagation.

To determine the extent of the improvement in reception made possible by this work, tests were carried out thru the worst summer months, namely, July and August, on various European stations, and it was found possible to receive the 6,000-meter Nauen signal some five or six hours per day during the worst periods when reception otherwise was totally impossible. Complete, continuous reception was not, however, yet possible, since there were times when, due to fading, the strength of Nauen's signal fell so low that it was no longer possible to receive it Very interesting, and surprising also, was the fact discovered thru the constant use of this arrangement, that the heavier the static disturbances were, the more perfect the balance which could be secured, and the greater the improvement in the static-to-signal ratio which resulted. This very significant observatin led to a careful study of the character of static disturbances under conditions of weak and strong disturbance, and it was noted that invariably the strong static consisted mostly of the grinders type, the percentage of this type increasing with the increase of total static energy and decreasing with the decrease of the total, and it may be said that the results of long and continued work since these first experiments has established the

facts referred to definitely and conclusively, the occasional variations therefrom being of such infrequent occurrence as to be negligible. It was also noted at this time somewhat unexpectedly, that the disturbances from a nearby thunderstorm were at times quite markedly reduced, the amount being seemingly dependent upon the position of the storm with reference to the receiving station. This improvement, however, was not of a sufficient order to render reception, thru local lightning, generally possible.

Many attempts were made to measure the improvement under various conditions in the signal-static ratio of this system, as compared to a single loop thru the use of the well-known audibility method, and the results obtained varied from more than a thousand times, under very severe conditions, down to five or ten times for very light static conditions. Now the audibility method measures the current in the telephone circuits from which it follows that the energies represented by two different audibility measurements are proportional to the square of the audibility factor; consequently the ratio of one thousand-to-one in audibility means one-million-to-one in energy. Unfortunately this method is a poor one for measuring static disturbances accurately and I cannot say that the above ratio is accurate. I find, from continuous use of this method of measurement, that, while it gives reasonably good results where the sound in the telephones is of a musical character, when this musical character is lacking the ear is unable to judge relative intensity accurately. In addition to this difficulty there is the fact that static disturbances are of extremely irregular intensity and that at any two successive instants widely different energies may exist. No other suitable method being available at the time, it was decided to depend on comparisons of readability of the signal resulting from the use of the complete system, as compared to the single loop, and this method has been used chiefly since that time.

A new method of measuring static intensities has recently been developed, which is the joint suggestion of Mr. G. H. Clark, expert radio aid of the Navy Department, the Research Department of the Marconi Company, and the writer—and which has been put into practical form by the Research Department. It measures the intensity of static disturbances in terms of the signal intensity necessary in order that the signal may be read, and it is hoped that a large number of measurements made by this method can be presented in a later paper.

Since the continued operation of systems of this type has so clearly emphasized the existence and characteristics of the two types of static referred to as grinders and clicks, a brief reference to some of the distinguishing characteristics may be of interest. It is found that the grinders type is most prevalent in the warm season, during warm days, and between the hours of noon and sunrise the following morning. The sounds which they produce in the telephone are generally a sort of continuous rattle, with occasional heavier crashes. This type behaves as tho vertically propagated, *and appears to affect antennas, separated considerable distances, simultaneously.* It can therefore be excluded, thru the use of the system described, while the signal is retained. The clicks, on the other hand, which sound like relatively widely spaced crashes, are most noticeable during the cooler periods of the year and day, but do not, except on very rare occasions, reach an intensity which is sufficient to interfere with the reception of signals, the strength of which is equal to the normal strength of Carnarvon or Nauen, or even Lyons. When the signal to be received, however, is of a lesser order of strength, such as that from Clifden, Ireland, or Eiffel Tower, or when the signals from the previous stations are abnormally weak, as occurs during sunset and sunrise fading periods, the intensity of this type of static is sufficient at times to cause great difficulty. It was found also that this type of static could not at that time be sufficiently reduced, thru the use of the system just described, to overcome the difficulty, and that adjustments which reduced it resulted also in a reduction of the signal. It appears probable, therefore, that this type of static follows the second of the two hypothesis previously given, and that it is in fact a true stray wandering in from all directions in haphazard fashion. How this vagrant was successfully dealt with will appear presently, but before getting to this point, which involves somewhat different arrangements than those shown, a brief reference will be made to certain modifications of the system with which experiments were conducted.

Midway between the loops of Figure 7 a third loop of similar dimensions and disposition was erected and used in conjunction with either of the two loops of Figure 7. This variation is shown in Figure 11, and it is to be noted that while the end loop has the long, horizontal lead, the middle loop has none, it being brought directly into the receiving station. It was with some interest that this arrangement was found to give rather better results than that secured with the loops separated the maximum distance

available, and the improvement was found to be in the perfection of balance, which was found to be sufficiently greater than that obtained with the arrangement of Figure 7 to more than offset the loss of signal on balance, due to the shorter spacing. This seemed to indicate, at first, that the farther apart the loops were the less perfectly could the static currents be balanced and the converse. Small loops of a large number of turns were

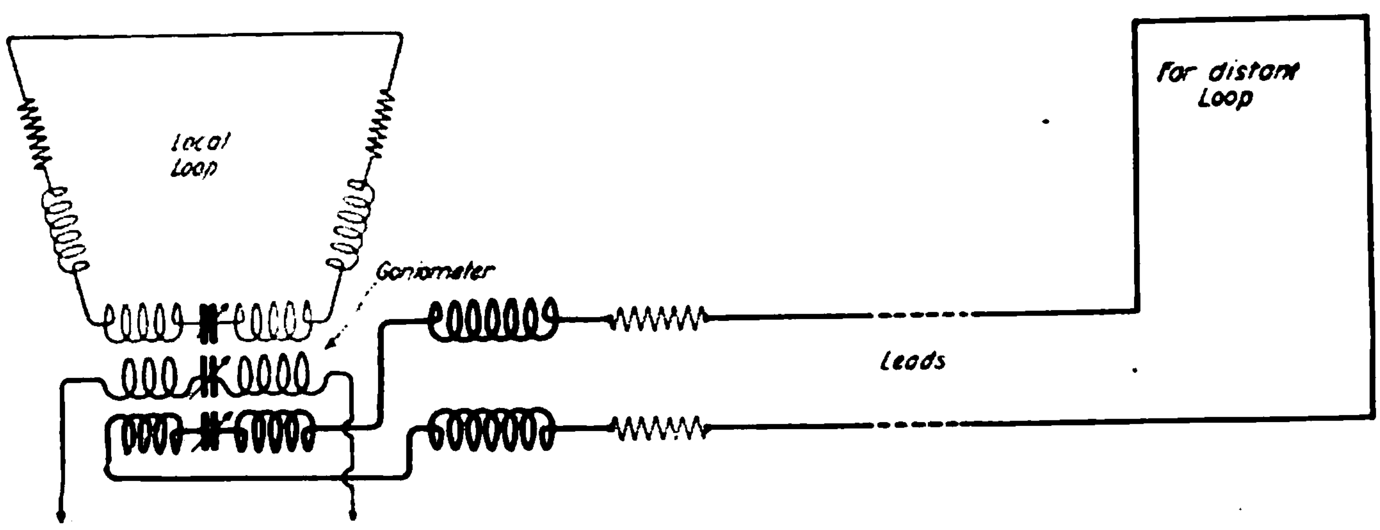

FIGURE 11

then erected at distances varying from ten feet (3.05 m.) up to 1,000 feet (305 m.), symmetrically located with respect to the receiving station, and in the same line with the big loops. Tests with these showed, however, no more perfect balance than that obtainable with the spacing of Figure 11 which was 2,500 feet (760 m.), and it therefore became evident that something besides the spacing accounted for the improvement in Figure 11 as compared with that of Figure 7. It would take too long to describe the very numerous experiments which were made in the attempt to run down this very elusive matter, but it was finally discovered that the reason for the performance above noted was the action of the long horizontal leads which, notwithstanding the fact that these were in the same horizontal plane and that the system had no earth connection, proved to be very effective aerials, picking up both signal and static. It was found that the static currents generated in them were in a definite direction and that consequently they must be connected to the loop in the same sense, that is, in such a way that the static currents generated in both the loop and the leads tended to flow in the same direction at any instant, so that when balancing at the receiving station all of the static currents generated in each half of the system were similarly affected. Before this fact was found out the arrangement of Figures 7 or 11 was so con-

nected that when adjustments were made which balanced out the static currents generated in the loop, those generated in the leads were added. The method of getting the right connection was simply to connect in a reversing switch, as shown in Figure 12 and to try the balance with the switches in each side of the system in the various possible positions. The best, of

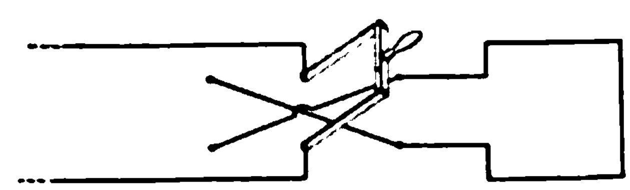

FIGURE 12

course, was that which gave the most perfect balance and was very easy to find. The results obtained thereafter were found to be better for the long than the short separation by an amount proportional to the separation, and it is interesting to note at this point that in all subsequent work the perfection of balance of static currents obtainable was the same, regardless of the overall length of the system which, as will appear shortly, has been in some instances as much as six miles (9.6 km.), while the signal combined always in proportion to the spacing.

In Figure 13 is shown an arrangement in which all three antennas were used. In this arrangement the two antennas at

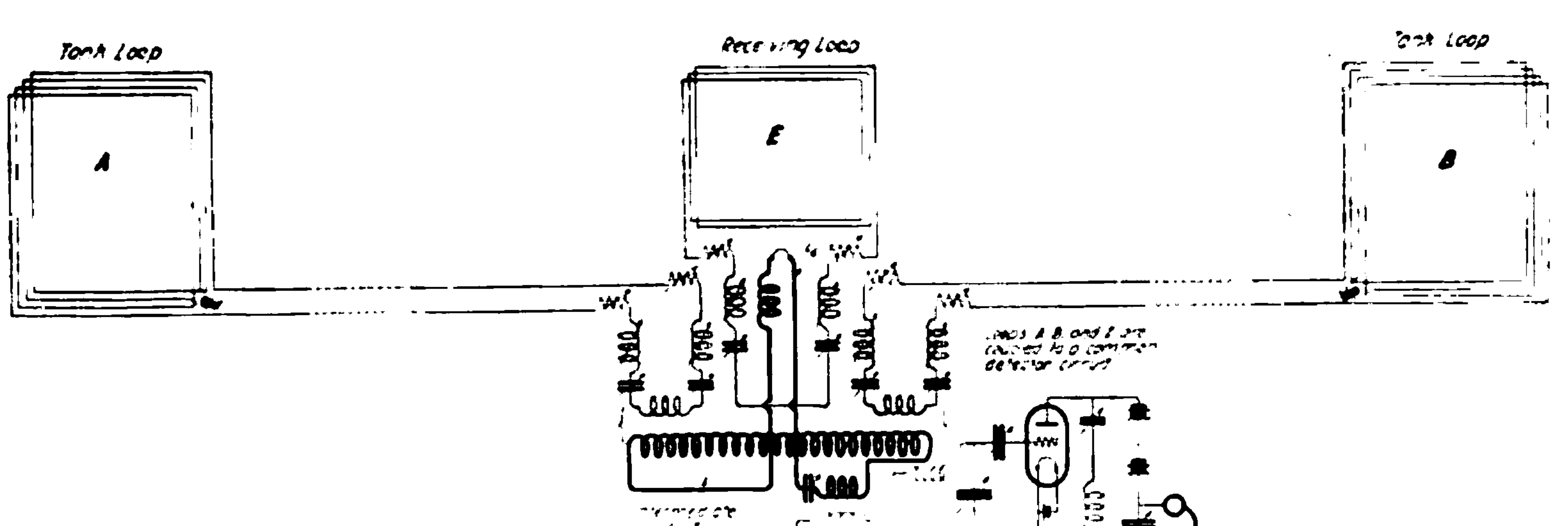

FIGURE 13

the ends were so coupled and adjusted that the signal was cancelled out, leaving most of the static. This arrangement has been termed the "static tank" since it was a source of static currents of any desired frequency without being a source of signal current. When this adjustment was accomplished the circuits

connected to these two antennas were opened and the third antenna connected in and tuned. This third antenna provided both signal and static currents in whatever ratio they happened to exist in this loop. The next procedure was to connect the two other loops in again and to adjust the intensity of the static currents from the middle antenna until they were equal to those due to the two end loops by use of suitable resistances, couplings, and so on. This third loop was connected into the system in such a way that the static currents due to it were opposed, leaving the signal due to the third loop. This arrangement resulted in a material improvement in working over those previously tried. From a consideration of the hypothesis of vertical propagation of static waves, it was not possible to account for this improvement, so far as the grinders are concerned, so that it was ultimately concluded that the improvement might be due to the elimination of some of the static of the other type. This possibility was somewhat supported by the fact that it is occasionally not possible to distinguish between the two types from the sounds which they make in the telephones since it happens occasionally that those of one type have the characteristic sound of the other. An analysis of the action of this system when affected by horizontally moving static waves, assumed to be uniformly distributed, brings out some most interesting facts.

Referring now to Figure 13, assume that the two aerials there shown have a spacing which is one-half the wave length of the signal received. Also assume that a static wave is arriving from the same direction and with the same velocity of propagation, and that this wave is so highly damped that no forced oscillation is produced in the aerial but that the only oscillation therein has a frequency and damping which is determined by the constants of the circuit. When this static wave, if it may be so called, arrives at the first aerial, an electromotive force is generated therein and currents start to flow thru it and the connected circuits. Current then begins to develop at the terminals of the detector. The wave continues its motion until it similarly affects the second loop and the resulting currents flow back to the receiving apparatus. Owing ,however, to the spacing which has been chosen, the currents from the first loop have had time to go thru a complete half cycle before that from the second arrives. Suppose now that the connections and position of the goniometer coil L_6 are such that the emf. produced in it by the signal is equal and opposite and therefore the signal cancels out.

Under these circumstances the results due to static will be as indicated in Figure 14, in which the solid line shows the damped oscillation due to the first loop, while the dotted line shows that due to the second loop when the method of connection is that just described. This diagram brings out the interesting fact

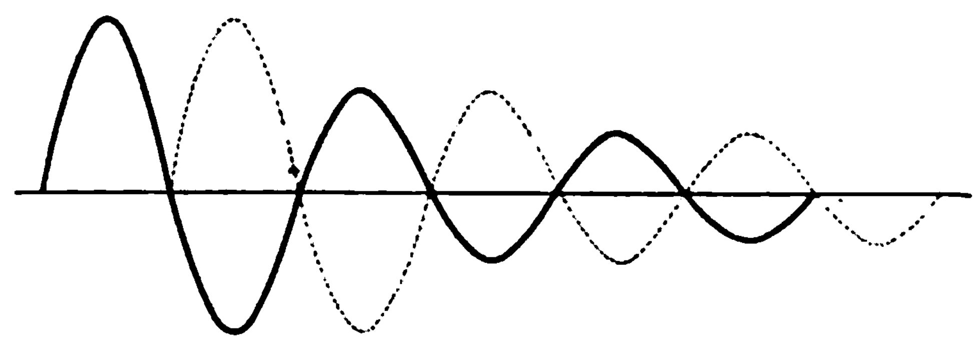

FIGURE 14

that the first half oscillation arriving from the first aerial is unopposed; that the first half oscillation from the second aerial is opposed to the second half oscillation from the first aerial which, because the oscillations are damped, is of smaller amplitude than the one opposing it. This condition obtains thruout the entire train so that the resultant of the two oscillations is not zero.

If now, the aerial circuits are heavily damped thru the addition of resistance, the wave train due to static becomes shorter and shorter, until when the limit is reached, all of the energy is in the first half swing. Therefore, under these conditions, while the circuits are so adjusted that the signal completely cancels out, yet the entire static current remains and we have the curious condition shown in Figure 15 of the two half oscillations, both in the same direction. It should be here noted that the first half oscillation of the signal is also unopposed, but if the signal current be undamped the percentage of the total signal energy which affects the detector is, roughly, 1/7,000th part of that which arrives during the time occupied by a dot at an ordinary rate of sending.

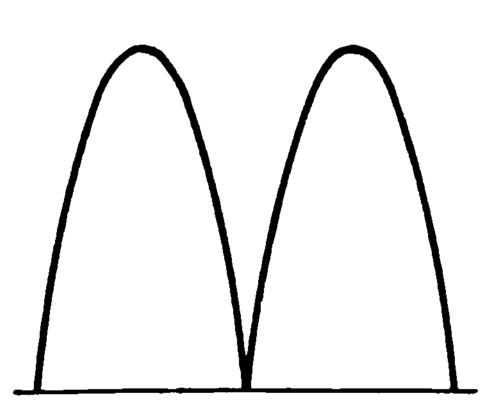

FIGURE 15

If we assume now that static disturbances are uniformly distributed thru all horizontal azimuthal angles, then as the angle, with the direction of the system increases, the intensity of these pulses decreases in proportion to the cosine of the angle; at the same time the effective phase difference between the loops decreases so that these pulses begin to overlap. It is also assumed that the intensity of the oscillation which these pulses can give rise to is proportional to the maximum ordinate as a first rough approximation. It follows then that as these pulses overlap, a distorted curve, as shown in Figure 16, results, but its effectiveness is not materially increased until the maximum ordinate of the resultant curve rises above that of the single pulse.

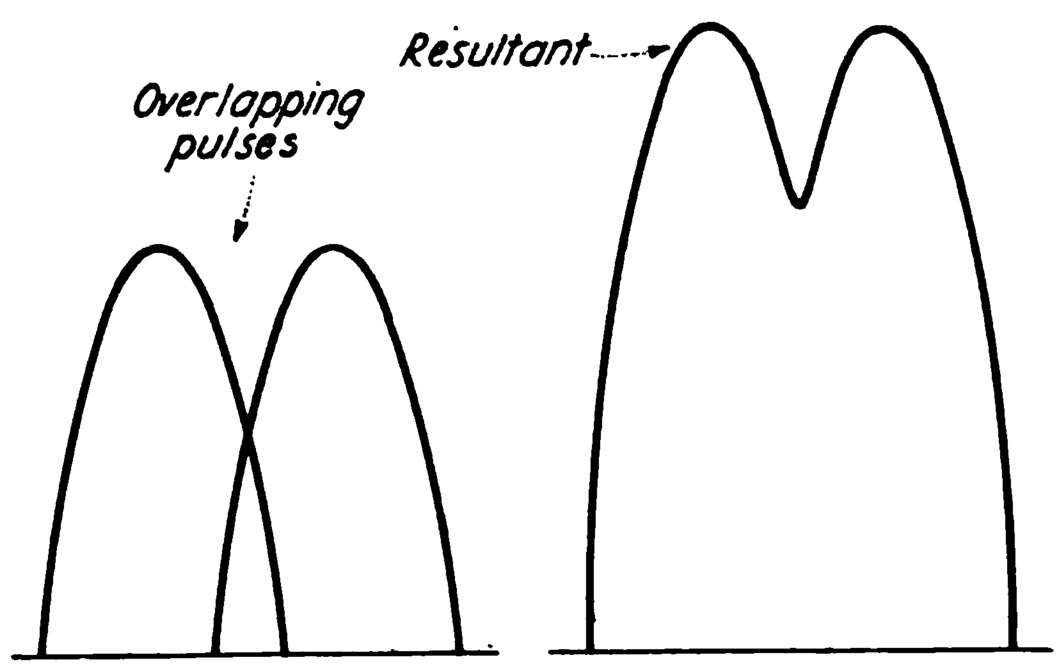

FIGURE 16

It will thus be seen that thru part of the azimuthal angle the intensity of the oscillations produced by static varies along the cosine curve, spreading out somewhat, however, as the angle increases. If, therefore, a third antenna is employed, the curve of reception of which is the cosine curve and in which both signal and static currents are flowing, and if this antenna be oppositely connected to the system just described, the static currents due to the click type of static will oppose and the residue will be of the order of the difference between the dotted curve shown in Figure 17 and the cosine curve shown in solid lines in the same Figure, from which it appears that a very large order of reduction is possible while at the same time utilizing the full signal strength developed by the third antenna. This explanation does not purport to be a rigorous analysis of the system described, but is presented merely as a rough approximation to the facts observed.

The arrangement just described realized these possibilities to an appreciable degree, but in the form then used was not capable of utilizing the possibilities above outlined to their fullest extent, and reference will be made later to another modification which displayed greater capabilities.

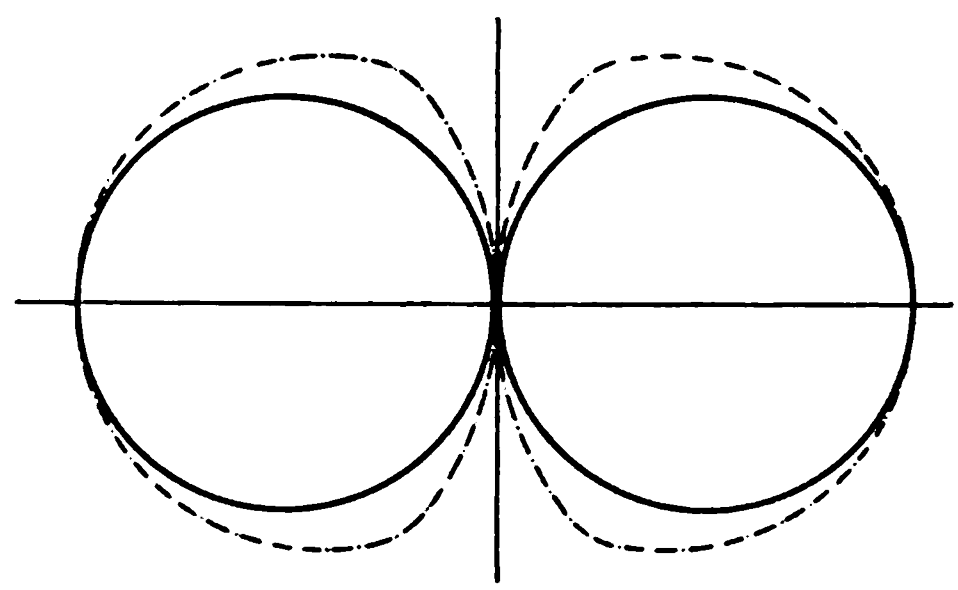

FIGURE 17

Various other combinations of the installation at Belmar were made. In Figure 18 the leads were disconnected from the loops and their ends joined, thus making of them horizontal aerials tuned to earth. It will be noted that this arrangement consists

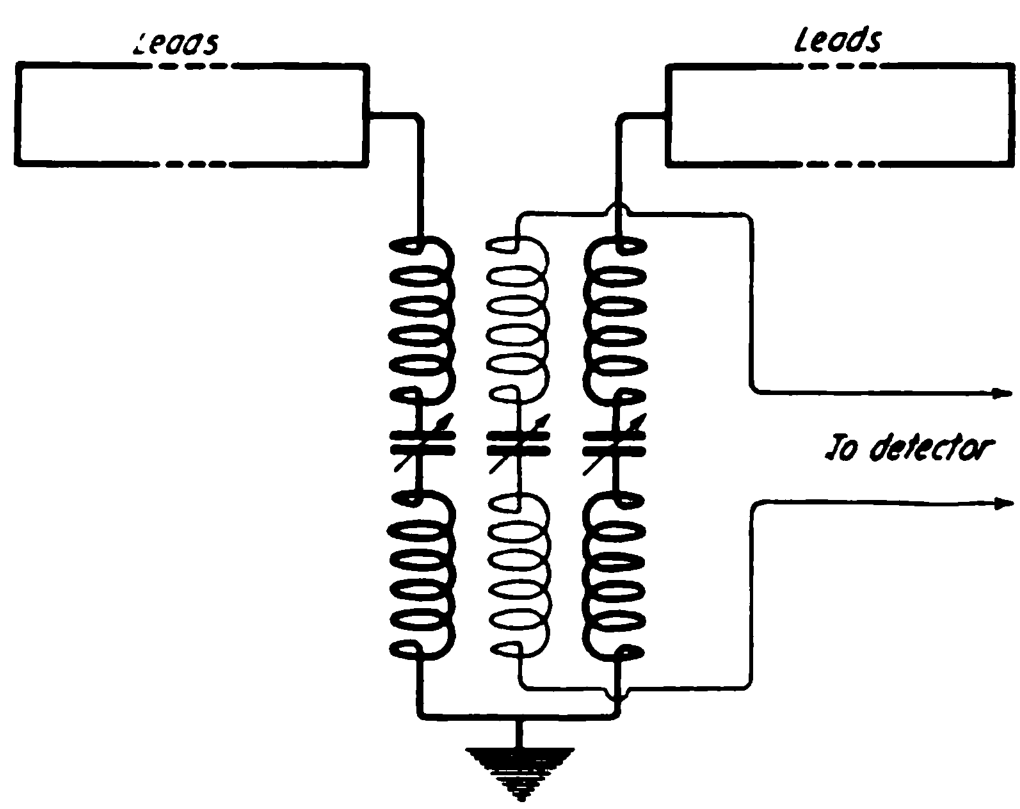

FIGURE 18

of two directive Marconi antennas and that the ratio of length to the height is unusually large. From this it follows that the aerial, which is pointed in a direction away from the transmitting station is a much better receiver of the signal energy than that

aerial which runs in a direction toward the transmitting station. Both aerials, however, pick up the same amount of static. From this it is evident that the two aerials may have a very marked difference in their signal-to-static ratio, and this effect will add to the effect resulting from their phase separation particularly when this separation is small, and constitutes at times a factor in the results obtained. This principle operates in all of the arrangements which will be described in which horizontal aerials are used, regardless of whether they are above the earth's surface, on the earth's surface, or underneath it. Figure 19 shows one of these in which the loop leads were connected together and each loop converted into an ordinary antenna tuned to earth. In

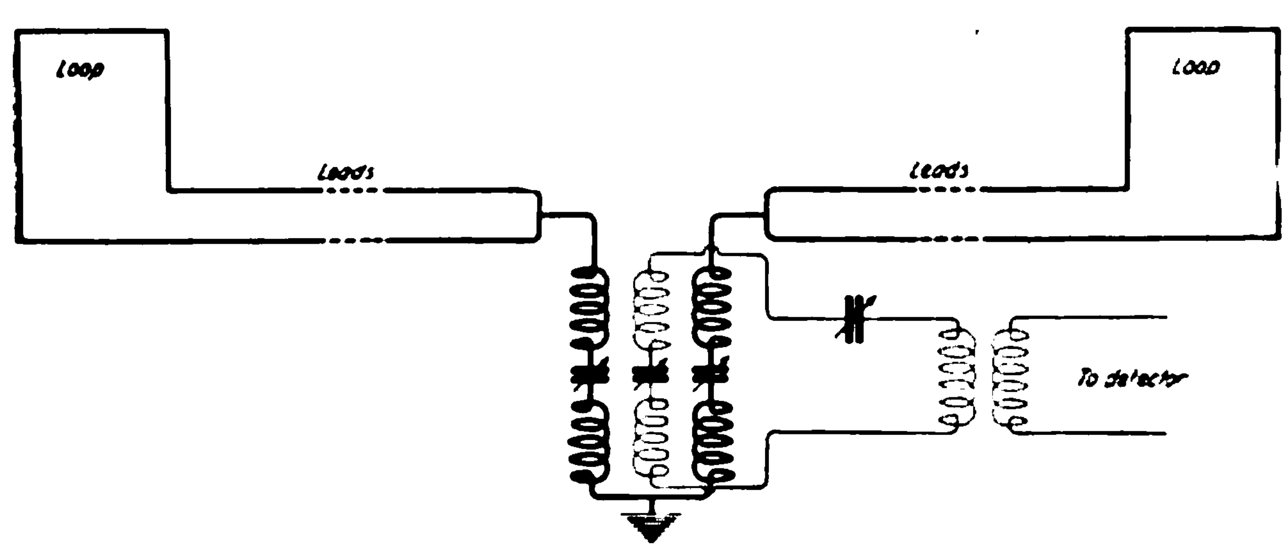

FIGURE 19

Figure 20 one loop was used in its normal way and balanced against the leads of the other loop tuned to earth. Figure 21 shows one loop connected in its normal way while the other one was arranged as an earthed antenna. All of these arrangements gave good results, but since it was impossible to investigate them all at once, the loop arrangements were chosen for first attention. Variations in the circuits were also tried. Figure 22 shows the parallel condenser arrangement, which was quite useful in se-

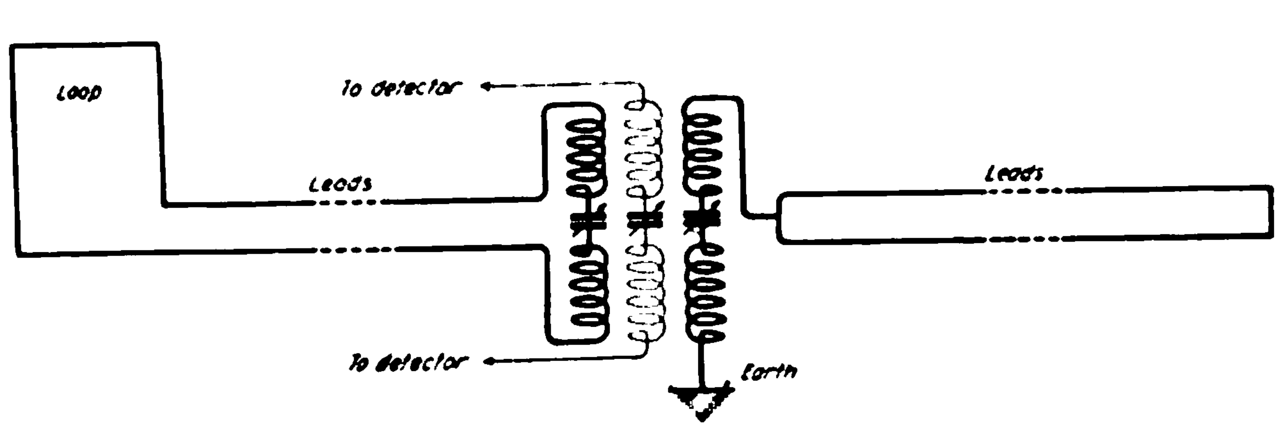

FIGURE 20

curing tuning to wave lengths shorter than could be obtained from the series condenser arrangement. Figure 23 shows another arrangement, in which most of the tuning was effected by condenser C_1 and inductance L_1 common to both circuits. Condensers C_2, C_3, C_4, and C_5, in addition to taking some part in the tuning, provided phase control and coupling.

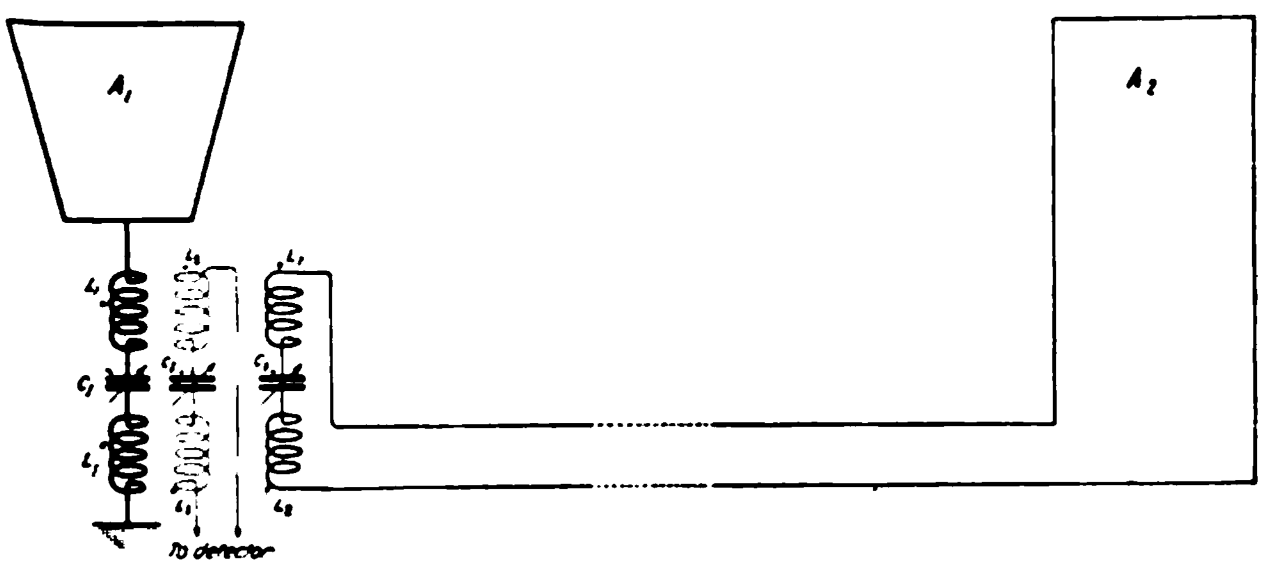

FIGURE 21

In addition to the capabilities of the arrangements described for receiving thru static, they have marked capabilities in working thru interference from other stations. When adjusted to

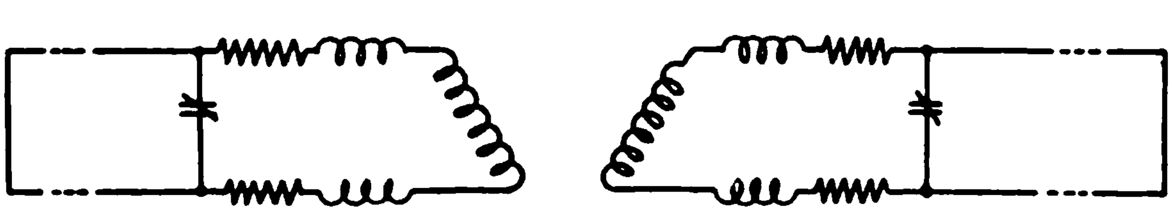

FIGURE 22

annul static of the grinders type this system has a reception curve of the form shown in Figure 9; its equation is $v = V \cos^2 \theta$, while that of the single loop is a cosine curve. It will be noted that the directional effect in this case is materially greater than with the single loop. When desired, adjustments can be so made

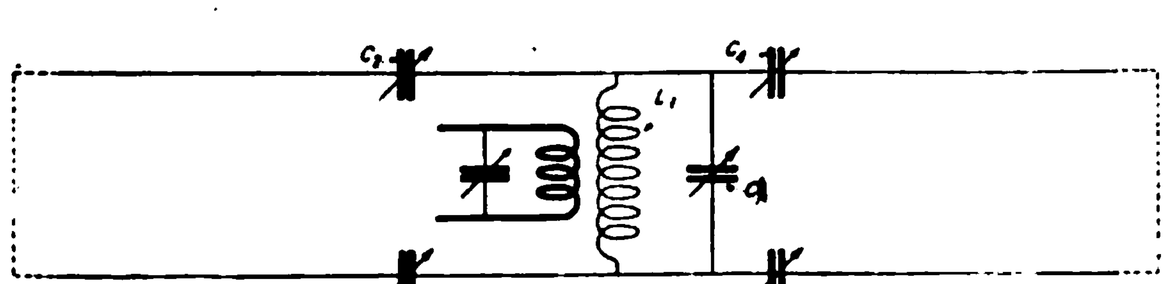

FIGURE 23

that the reception curve becomes that of Figure 10. Between $\theta = 0$ and $\theta = \pi$, the curve is a (cosine)^2 curve while between the angles $\theta = \pi$ and $\theta = 2\pi$, the curve is a (sine) $\cdot$ (cosine) curve when the loops are one-quarter wave length apart. This curve indicates that while reception in one direction is a maximum, reception from the opposite direction is zero, while it is also materially reduced in the third and fourth quadrants. This line of zero reception can be swung around at will thru the third and fourth quadrants by alteration of the phases of the currents in the two loops so that interference from any station arriving in this quadrant can be annulled, while reception is maintained from signals arriving in the first and second quadrants. It is to be noted that advantage can be taken of this property to eliminate strays, if they happen to be coming from a direction other than that from which the signal arrives, and this fact is of great help when a thunderstorm is gathering in the vicinity of the station. The necessary method of adjustment is as follows:

Suppose that the two loops of the system are one-quarter wave length apart and that the desired signal arrives from right to left. Then the currents in the left-hand loop are 90 degrees behind those of the right-hand loop, if the circuits are accurately tuned, and they will add in quadrature. Next, suppose a signal arrives from left to right; then the currents due to this signal in the left-hand loop are 90 degrees ahead of those in the right-hand loop and therefore also combine in quadrature. Then currents due to both signals exist in the common receiving circuit.

Suppose now, the phases of all currents in the left-hand loop are shifted forward 90 degrees; then the currents due to the desired signal in this loop are shifted around until they are in phase with those from the right-hand loop, while the phase of the currents due to the interfering signal in this loop, and which were previously 90 degrees ahead of those due to the right-hand loop, are now 180 degrees ahead of those in the right-hand loop, so that they oppose and neutralize. Because of the unusual characteristics of the antenna used, this shift in phase is readily accomplished by a small adjustment of the condenser in the loop circuit. If the interfering signal is not in line, the phase shifting can be made the right amount to take care of it, and this general order of result is obtainable to some extent with any spacing between the loops, although one-quarter wave length is best. The reception of Carnarvon's signal, 14,200 meters, thru the powerful interference of the 200-kilowatt

Alexanderson alternator at New Brunswick, only 25 miles (40 km.) away, working at 13,600 meters, has been an everyday performance of the system, while at the same time preserving a good static balance. All forms of the arrangement described have capabilities of reception thru interference, these capabilities varying with the type of antenna employed, the loop antennas and the horizontal aerials giving similar curves.

Work with the original installation soon indicated the desirability of increasing the spacing between loops both to secure greater signal intensity and also to determine whether or not the static currents would be simultaneously generated in the two aerials, when the spacing was thus increased. Antennas were therefore erected approximately 8,000 feet (2,430 m.) apart, each antenna consisting of 12 turns approximately 77 feet (23 m.) long by 30 feet (9.2 m.) high, supported from cross-arms attached to telephone poles. This construction is shown in Figure 24. The receiving apparatus was located at a point near the

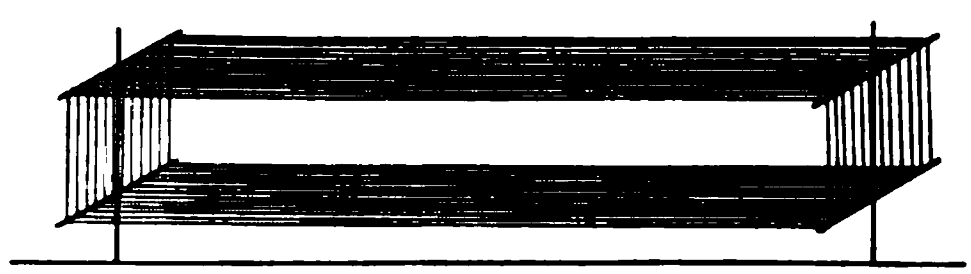

FIGURE 24

northeast loop instead of in the middle, and leads similar to those of the original arrangement run out to the southwest loop. When this was tried it was found that the leads picked up more signal and static than the loops and that the intensity of all currents from the southwest loop was so much greater than that from the northeast that successful working could not be obtained. Leads running along the ground, spaced at various distances, were then tried, and it was found that their effect was a minimum when the leads were close together. Next, a duplex lead-covered cable was tried and the effect of the leads very greatly reduced thereby. These leads of course had enormous capacity for a circuit of this sort with the result that the southwest loop was connected to the receiving station thru a capacity coupling of very small value, and in order to get equal signal from the distant end it was necessary to use four similar loops at that point connected in series-parallel. It was also found necessary to have a tuning condenser and inductances,

shown in Figure 25, located at the remote loop and an operator stationed there to make adjustments in accordance with instructions telephoned to him by the observer in the receiving station, using the cable wire for this purpose. It was also neces-

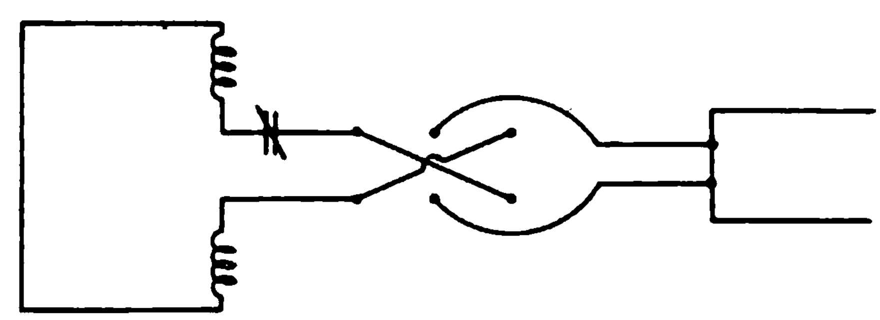

FIGURE 25

sary to use the reversing switch previously described, since even the lead-covered cable picked up signal and static in appreciable amounts.

By the time this work was completed the season had advanced well into the winter and the amount of static available for working was so small that the results obtained were inconclusive and the work was abandoned for the time being.

In view of the entrance of the United States into the war and the great national importance of the improvements previously described, a full disclosure of the most fully developed two-loop arrangement was made to the Navy Department and an official test carried out by Mr. G. H. Clark, expert radio aid of the Bureau of Steam Engineering of the Navy. The following is a quotation from the report received by the Marconi Company from the Bureau relative to this test:

"The Weagant circuit in its present form will enable trans-Atlantic reception to be carried on without interruption in so far as elimination of static is concerned."

The next experiments were conducted at Miami, Florida, where loops were arranged at varying distances, the maximum being six miles (9.6 km.) and the minimum about 100 feet (30.5 m.). Having in mind the difficulties due to the leads, a special lead construction was used in which a pair of number 18 wires*, spaced about two inches (5.08 cm.) apart, were run thru pasteboard tubes about three inches (7.62 cm.) in diameter, these tubes being in short lengths joined together and covered on the outside with tinfoil. It was thought that this arrangement would give a reasonable value of capacity between the leads

* Diameter of number 18 wire = 0.041 inch = 0.16 cm.

while the tinfoil covering might act as a screen in preventing signal and static currents from being picked up by the leads. This latter result was a desirable one since the greater the extent to which the leads act as aerials, the shorter is the effective spacing between the two aerials for a given total length. The results obtained with this lead construction were slightly better than those obtainable with any other, but the improvement over the results secured from the use of two similar wires, similarly spaced, but not surrounded with a shield, was of too small an order to warrant the expense and trouble of the other type of construction. Two loops of the type shown in Figure 24 but 150 feet (46 m.) in length and three miles (4.8 km.) apart, were connected to a receiving station located midway between them, and tests were conducted with various European stations, and it was found that the balance of static currents secured was as good as that obtainable with loops a short distance apart, while the signal strength at balance was much greater. This arrangement was not, however, generally satisfactory, as the loops were not large enough to give a satisfactory intensity of signal for practical working, while the effect from the leads was about equal to that from the loop. Two other loops were therefore constructed 7,200 feet (220 m.) apart, of the same general type and height, but of twice the horizontal length, and with these two, very satisfactory practical working was secured. With both of these arrangements the local tuning at the loops previously described was necessary, and this always involved a tedious adjustment until the correct setting for a given wave length was obtained, and even when this setting was known, it was necessary for some one to go to each of the loops,—not a convenient procedure with antennas three miles (2.2 km.) apart.

In order to overcome the objection just mentioned, the arrangement of Figure 26, which was the joint suggestion of Mr. Frank N. Waterman, and the writer, was constructed. As is indicated by the Figure, each loop consisted of a single turn extending from the station out and back again, thus being both loop and lead simultaneously, and being free from points where abrupt changes in circuit constants take place, as in the previous arrangement. The loops of this form which were constructed varied in length from 1,000 feet (305 m.) each up to approximately 9,000 feet (2,750 m.) each, the upper wire being supported on stakes only three feet (92 cm.) above ground, while the lower wire ran along the surface of the ground. Much difficulty was experienced in maintaining this construction long enough to

get satisfactory observation, due to the fact that about 2,500 feet (760 m.) northeast from the receiving station, they had to cross a canal, while at other places they ran thru cow pastures and were frequently broken.

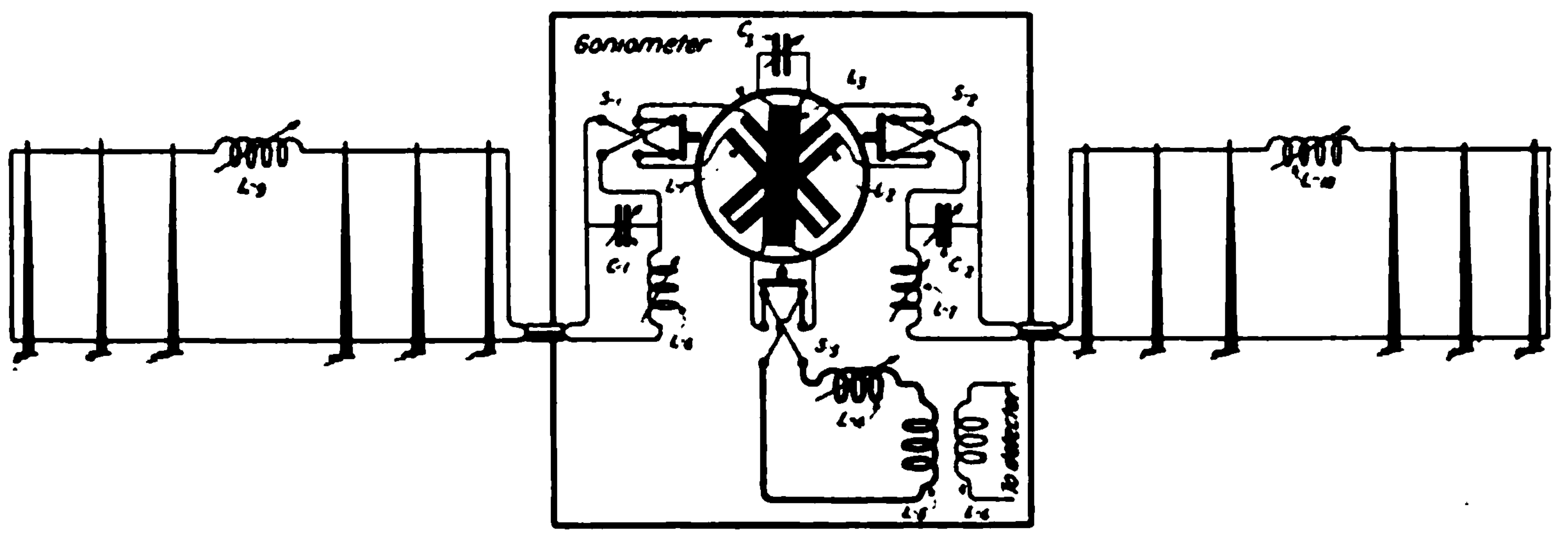

FIGURE 26

When first tested it was found that the loop then used, which was approximately 3,600 feet (1,100 m.) long, would not tune, the inductance and capacity inserted at the receiving end of the station apparently having no effect. This result was believed to be due to a current distribution in the loops of such a nature that there was a current node at the point of insertion of the tuning devices. It was therefore determined to attempt to alter this distribution by the insertion of inductance at some suitable point. An inductance such as L_9, L_{10} in Figure 26 of 30 millihenrys was inserted successively in the upper wire at a large number of points between the receiving station and the other end of the loop. It was found that the tuning improved constantly as the coil was moved from one end toward the middle, and constantly became poorer as the coil moved from the middle toward the end, the curve of the resulting effect being of the form shown in Figure 27. Insertion of the inductance in the lower wire produced no result and in fact if inserted in the middle point of the lower wire at the same time that inductance were inserted in the middle point of the upper wire, the effect of the latter was annulled. Having determined the best point for the inductance, its best value was next obtained, and while the results showed that a value of 30 millihenrys was about right for a wave length of 12,000 meters, and 5 millihenrys for a wave length of 6,000 meters, either value was sufficiently acceptable for both wave lengths.

As soon as tuning control of this type of antenna was ac-

complished it was found possible to use this system in a most satisfactory way for the elimination of static. The effective spacing of two such loops was found to be approximately the distance between the centers and complete control could be

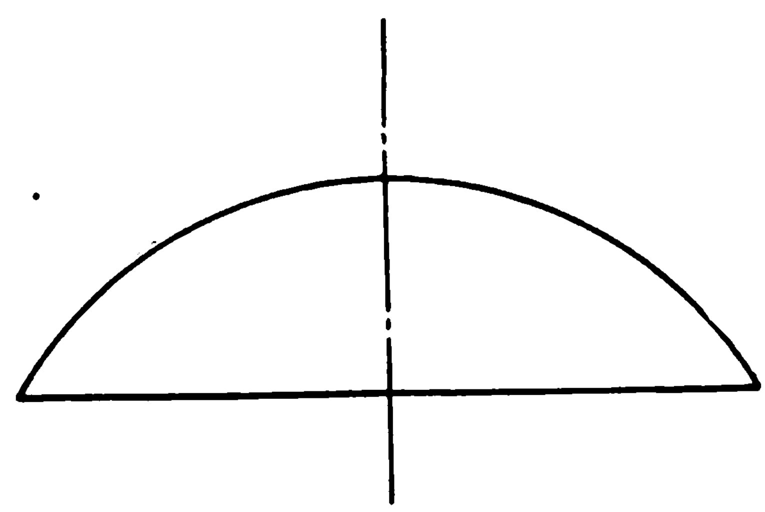

FIGURE 27

effected at the central receiving station, the over-all results obtained being even better than those obtained with the previous forms. A variation in the form of this type of aerial is shown in Figure 28, in which the area enclosed is approximately a triangle, it being assumed that this arrangement would give a greater effective separation if the loop receiving antenna extracted energy from a passing electro-magnetic wave in accordance with the usually accepted theory. Conclusive results on this form were

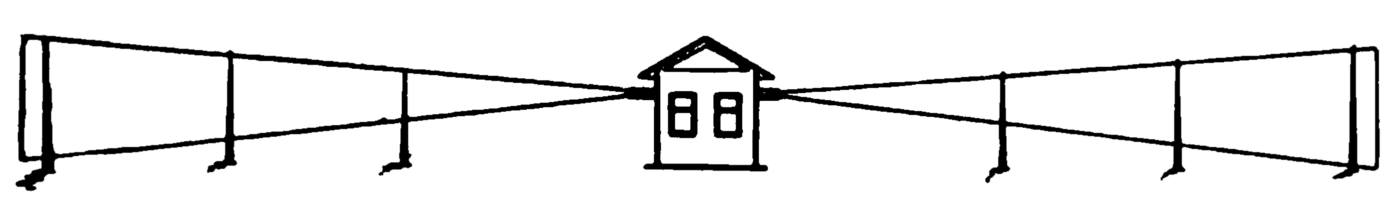

FIGURE 28

not obtained until later work at Lakewood, New Jersey, and they showed that the very long triangle there used, *did not* behave in accordance with this assumption. In fact it may be stated that this whole work has demonstrated that our ideas of the mechanism by which a loop antenna extracts energy from a moving electro-magnetic wave, will have to be considerably modified, but this matter is too extensive to go into in detail at this time. The exact mode of vibration of the long, low loops just described is also a matter of great complexity and can only be determined by an exhaustive experimental and mathe-

matical analysis. This work is already quite well under way, Mr. Louis Cohen having kindly consented to undertake the mathematical work for me, but it is not yet completed.

During the early part of the work at Miami, the Navy Department was experimenting with underground antennas, several of which had been installed by Mr. G. H. Clark. This afforded an opportunity for the writer to try these in the system described. These were tried in a large number of combinations, which are shown in Figure 29. As will be seen from the Figure, these arrangements were essentially the same as those of the original Belmar installation, resembling most closely that one which employed the horizontal lead wires and differing from it only in the fact that the wires were buried under ground or laid on the surface instead of supported some distance above the ground. All of these various combinations operated satisfactorily, but there was no material difference between those laid on the ground, those under ground, or those under water, which was present only a few feet under the surface. The lengths used were not great enough to give a material fraction of a wave length effective spacing, and attempts were made to extend this by increasing the length of the wire. It was found, however, that this could not be done, but that, on the contrary, increasing the length of the wire made the performance poorer rather than better, and this is probably due to the loop action and other causes referred to in the preliminary description of this type of aerial. The best working of all these arrangements was the combination of one of the ground wires with one of the loops, due to the fact that this gave a much greater effective separation, the loop being situated 3,600 feet (1,100 m.) away from the receiving station.

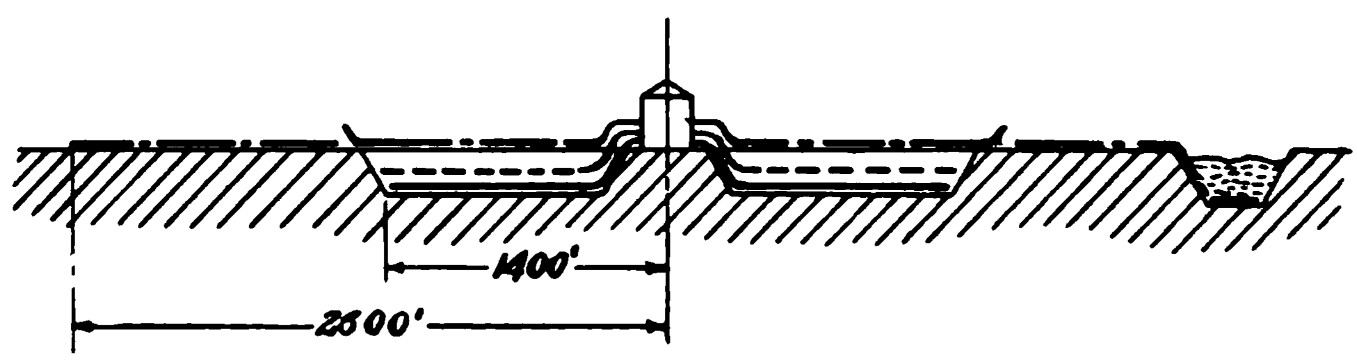

FIGURE 29

Having secured a practical form of this system which could be operated with the half-wave-length spacing, namely, the long, low loop, an installation was made in the spring of 1918,

in the vicinity of Lakewood, New Jersey, in accordance with the results of the Miami tests. This installation consisted of two aerials, each three miles long, of number 14 hard drawn copper*, in a line directed toward France. These antennas were supported by telephone posts 30 feet (9.2 m.) high and were at first triangular in form, having a vertical leg 28 feet (8.5 m.) high at their outer ends, and brought together at the receiving station. This is shown in Figure 28. This form was later modified to a rectangle three miles (4.8 km.) long, ten feet (3.05 m.) in vertical dimension, the lower wire being about ten feet (3.05 m.) above ground, and this modification was found to be appreciably more satisfactory. Inductance coils of 30 millihenrys were inserted in the middle points of the upper wire of each loop. This station was operated continuously from the middle of July until the end of September with a force of three operators, each working eight hours, copying messages sent out by Lyons, Carnarvon, and Nauen regularly, and occasionally other stations. This continued operation was undertaken to determine the capabilities of the system in a practical, commercial way, during the worst period of the summer and at all hours of the day. The results secured were most gratifying, the total interruptions experienced being of no greater total duration than those of good cable working between the same points and at the same time of year. It was found that when the signal from the European stations was of normal intensity the heaviest static experienced at any time was unable to interfere in the slightest, but that on the contrary it might have been very much more severe without causing trouble. Reception under this condition was almost invariably good enough for high speed automatic reception. A few thunder storms occurred during this time and some, but not all of them, prevented reception while they lasted. There were also periods recurring regularly every day between four and six o'clock in the afternoon and between twelve and two o'clock in the morning when the intensity of the received signals from Carnarvon and Nauen fell off enormously, on some occasions falling as low as 1/100th of their normal intensity. During a few of these fading periods interruptions were experienced varying from five or ten minutes to perhaps one hour. The worst of these periods was usually, but not always, the midnight-to-two-a.m. period when, altho the static was generally lighter than during the afternoon fading period, at which time its maximum intensity occurred, the decrease of signal strength was rather greater.

* Diameter of number 14 wire = 0.064 inch = 0.162 cm.

A careful study of the conditions during these fading periods convinced me that the difficulty was due to the fact that when the signal weakened greatly the click type of static was present in sufficient quantities to cause the trouble, and that when the signal intensity was greatly amplified in order to be heard, the amplification also brought up this disturbance with its ratio to the signal unaltered.

As has already been pointed out, the two-aerial arrangement has not as yet shown itself capable of sufficiently differentiating between this horizontally moving type of static and the signal to meet the severest conditions of signal fading. In the hope of successfully overcoming even this condition, existing occasionally during the fading periods, recourse was had to the three-aerial arrangement, but in a modified form, as shown in Figure 30.

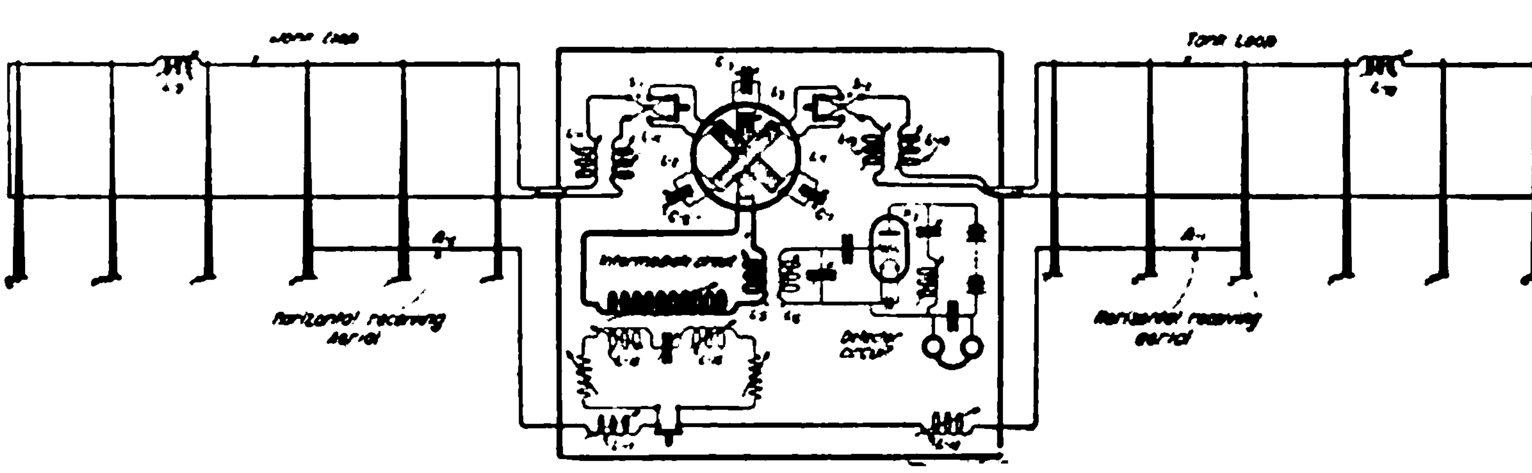

FIGURE 30

This was operated in accordance with the principle previously set forth, namely, the two loops were adjusted to balance out signal instead of static, and the retained static was used to balance out that in the third antenna. The third antenna in this case was a horizontal wire approximately 6,000 feet (1,820 m.) long, about three feet (92 cm.) above the ground, running underneath the loop antennas and supported by the same poles. When this arrangement was operated it was found that all of the hoped-for improvement, and much more, had been realized, in fact that the improvement of reception thru static of the stray or click type was of the same order as the improvement which the two-antenna arrangement made possible thru static of the grinders type; also, most fortunately, that the adjustment which reduced one type of static was the exact adjustment for eliminating the other, so that both types went out together, leaving all of the signal supplied by the third antenna. This was approximately the same in strength as that which could be received

using the two loops so connected that their signal strengths added.

So great was the general improvement in reception made possible by this arrangement that signals from stations in Europe of a very much smaller order of power than Carnarvon or Nauen could then be received. Of these stations it is sufficient to mention Eiffel Tower, working at about 8,000 meters, and Lyons, working at 8,000 meters, the signal strength of Lyons at this wave length being very much less than the signal received from the same station when using his usual wave length of 15,000 meters, and it is assumed, though not definitely known, that the amount of power being used was much less. The installation at Eiffel Tower is understood to be an arc, the input of which is about 100 kilowatts. Many attempts had been made during the summer to copy these stations with the two-antenna arrangement, but the results were satisfactory only occasionally and when the grinders type of static was that which existed. When the other type was present these stations could not be read. During the test with the three-antenna arrangement on one occasion, in the evening, static of extreme intensity was experienced and the intensity of the signal from Eiffel Tower was much below normal, with the result that with the two-antenna arrangement it was barely possible to tell that the signal was present. Using the three-antenna arrangement the signal was not only readable but of such intensity that it could be read with the telephones a couple of feet from the ear. Continued use has established beyond question that this performance is not occasional or accidental, but consistent, and that with this arrangement trans-Atlantic radio telegraphy can now be carried on free from interruptions due to static of any kind whatsoever except local lightning. This cannot always be neutralized, but since the cables are also interrupted by this latter cause it follows that a continuity of communication equal to that of cable operation is now possible by radio telegraphy, while the latter has the great advantages of cheapness and greater speed of operation. For many years, attempts to work automatic high-speed radio telegraphy have been made, but they have been successful only when static was absent. It is therefore evident that use can now be made of this method of working to a very great extent, thereby greatly increasing the number of messages which can be handled over a given circuit. It may also be stated that the great barrier in the way of successful, practical radio telephony has been removed since static has interfered with radio telephony to a much greater extent even than with radio telegraphy.

One of the outstanding features of the systems thus far described has been their need of a considerable stretch of territory, and it would obviously be an important advance if the same results could be secured without this necessity. It is pleasing, therefore, to be able to state that a considerable number of such arrangements have been worked out, in which the necessity for large space does not exist; in fact some of them are of such small dimensions that the entire equipment necessary, including the antennas, could be arranged in a lecture room, and between the floor and the ceiling. Only one of these arrangements will be described at this time, the others being reserved for a later communication.

Referring now to Figure 31, A_1 represents an aerial of the linear type several times referred to, but so arranged that it can be moved thru a considerable angle in the vertical plane and swung around thru any desired azimuthal angle. If this aerial

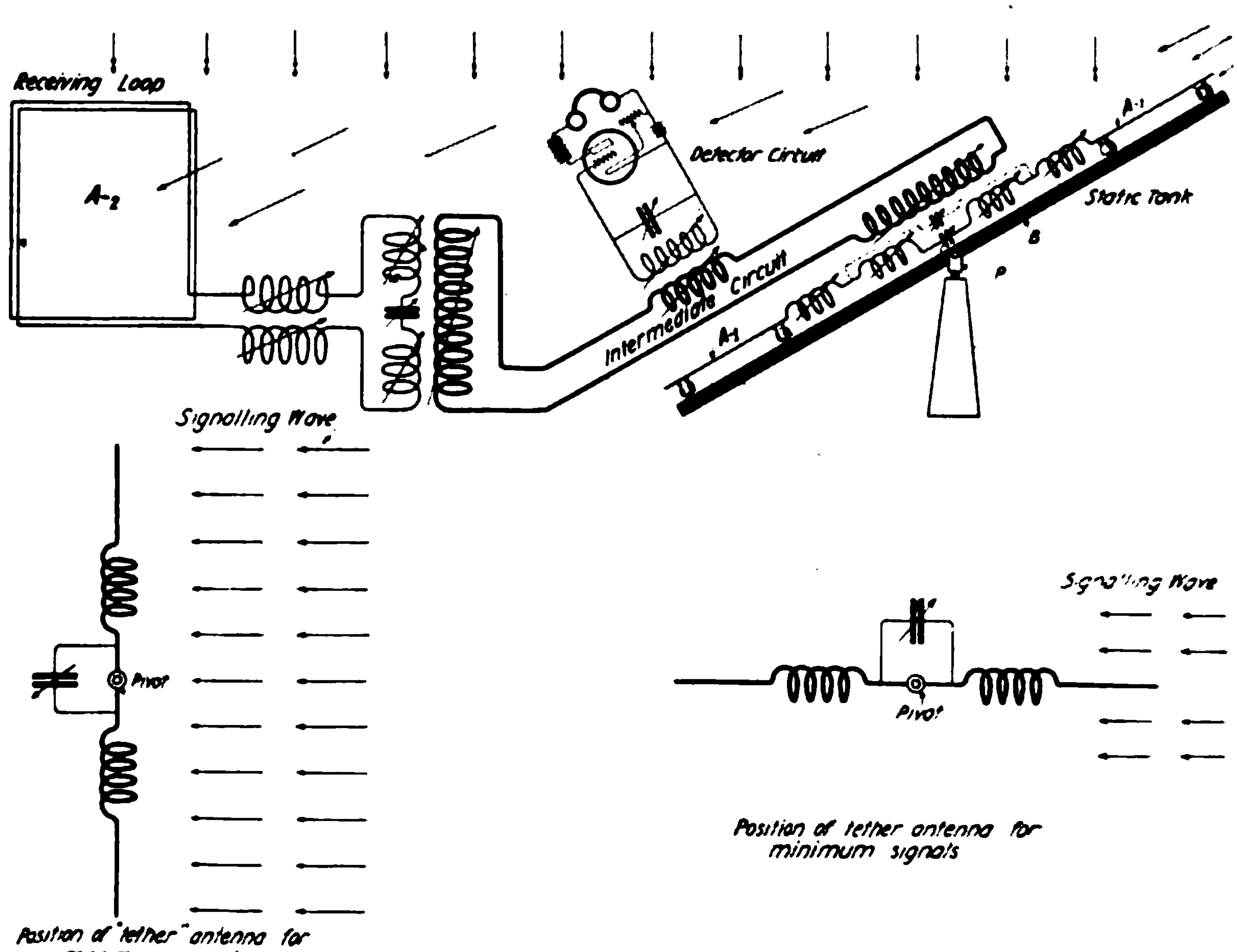

FIGURE 31

is swung from the vertical position to the horizontal position, while being directed toward a desired transmitting station, it is found that a particular vertical angle can be obtained at which the signal goes out entirely while some of the static re-

mains. The arrangement therefore constitutes another form of static tank. If then we take a second antenna, such as the loop A_2 shown in the Figure, which supplies both signal and static, we can, thru the use of the circuits already explained, couple these together in such a way that the static is cancelled, leaving the signal. This arrangement works quite as well as any of those previously described, except the three-antenna arrangement, and the difference is due to the fact that this particular arrangement does not as completely eliminate the static of the horizontally moving type.

To secure the full practical benefit of this arrangement it is desirable that the length of the aerial A_1 be conveniently short, say about 30 feet (9.2 m.), which, for trans-Atlantic reception of course makes necessary the employment of amplifiers of extraordinary capabilities. In the work which I have done with this arrangement I have used two amplifiers, developed by the Research Department of the Marconi Company, of five and eight stages respectively, with which it is possible to receive signals of a satisfactory strength from Nauen or Carnarvon when the antenna is of the dimensions stated. The reason for the use of the very small antenna is simply that it is then possible to secure a conveniently operated mechanical support, the principle of operation holding good, however, when much greater lengths are employed.

No attempt has been made in this paper to set forth exhaustively the complete theories of the arrangements used, but simply to give a brief description of the methods used and results obtained .It is realized that quantitative data of many sorts have not been given, due in part to the unsatisfactory nature of the methods of measurement available, and also to the fact that the whole work was dominated by the practical requirement of securing readability of trans-Atlantic signals, which was of vital consequence to the Marconi Company. It is hoped, in subsequent communications, to remedy the deficiencies referred to and to describe a considerable number of other arrangements which have been discovered, the operating characteristics of which are being more fully investigated.

With reference to the hypothesis stated in the early part of this paper to the effect that static of the grinders type is due to electro-magnetic waves heterogeneously polarized and propagated in a direction perpendicular to the earth's surface, it should be clearly understood that while this hypothesis has been of great use in explaining the very large number of observations

made over a long period of time, it cannot however be regarded in the light of a proven theory, since certain observations have been made which it does not readily cover. It does appear, however, that the most important fact contained in this hypothesis, namely, that static effects of the grinders type are simultaneously produced at points separated a distance of the order of the longest waves at present employed in radio telegraphy, is firmly established. With respect to the rest of the hypothesis, further work, which is at present in progress, will be necessary before the complete truth is proven or otherwise.

The writer would greatly appreciate the opportunity to demonstrate the actual working of the systems described thru a committee to be selected by THE INSTITUTE OF RADIO ENGINEERS should they be sufficiently interested, during the coming summer, when static disturbances are at their worst.

In conclusion the helpful and valuable assistance of the following gentlemen is acknowledged: Mr. C. L. Farrand, Mr. Frank N. Waterman, Mr. George H. Clark, Dr. Alfred N. Goldsmith, Mr. Louis Cohen; and of the Research Department of the Marconi Company. Messrs. Weinberger and Dreher of the Research Department deserve special mention in this latter connection.

SUMMARY: The effects produced by static (strays) are considered, and the Eccles classification of static as grinders, clicks, and hisses is adopted. Previous attempts to eliminate strays are described with explanations of their non-operativeness.

Researches are described which indicate that grinders, the predominantly objectionable summer static, act as if propagated vertically. A balanced antenna structure consisting in effect of two horizontal loops is used to eliminate grinders. Signals in the loops add in the secondary circuit in proportion to the separation of the loops relative to the wave length, while vertically propagated grinders in the loops balance out in the secondary.

Clicks are found to be horizontally propagated strays, and a special three-antenna arrangement for practically eliminating them is described and explained.

Experimental work at various stations using these arrangements is described, and the capabilities of the new systems indicated.

DISCUSSION

Michael I. Pupin: I have been very much interested in the paper, since it is a report of work actually done. Perhaps one does not care so much about the theory which has been advanced to explain the balancing described; whether the theory is correct or not, the effects produced show that it is a theory not so very far removed from the truth, if it is not exactly the true theory.

What strikes one most in listening to the paper is that we have here a new attempt to advance the art of radio telephony and radio telegraphy, an attempt, namely, in the direction of adjusting phases. In the beginning of the art, we had pulse excitation and nothing else; then came loose coupling and less damping, which made tuning available; it is still holding its own in the art; and then came attempts to produce continuous oscillations, and now we have a new addition to the art: namely, the art of adjustment of phases in the receiving antennas.

I myself am a great believer in this new departure. I have been its advocate for some time; namely, the adjustment of phases, and as we are dealing in radio telegraphy and telephony with wave propagation, it goes without saying that in wave propagation we can accomplish a great many effects by the proper treatment of phases, a feature which is not possible in ordinary transmission, which does not depend upon wave propagation.

The effects produced by loops pointing in the direction of the station which transmits, properly adjusted as to their length, and provided at the point of symmetry with a suitable receiving apparatus, can be adjusted so as to get a good signal, because you arrange the phases in a suitable way, which at the same time can be made to neutralize the static effect, if that static effect is not propagated in the same way as the signal. I say this seems to be obvious—*when you are told about it*, and that is what I like about the scheme. A scheme, which appeals to you as obvious after you are told about it, is the right scheme, as a rule.

It is very easily understood, and everyone will say "Why, of course, that will work." But it is rather difficult to understand why an electric wave due to the propagation of a grinding "stray" or a grinding static should affect a long conductor, several thousand feet long, in such a way that the motions of electricity in all its parts are affected in the same phase. It is very difficult to understand.

Mr. Weagant explains it by assuming that the wave is propagated vertically downward, and perpendicularly to the surface of the earth, as the wave is due to heterogeneously polarized oscillators—I think that is what they are called. Well, these oscillators must be not only heterogeneously polarized, but they must be somewhere above, because if they are at a distance, say one thousand miles (1,600 km.) away, or even five hundred or three hundred or even two hundred miles (800 or 400 or 300 km.) away, then their waves would not affect all parts of a long antenna alike. It is not easily seen how these oscillators start waves (which after awhile must become spherical) which, as far as the receiver is concerned, appear as if they came from above.

That is the difficulty. What of it? The more difficult the thing appears, the more interesting it is. It may be that these oscillators throw out spherical waves which are reflected back and forth between the conducting upper layer of the atmosphere and the earth, an assumption which has been advanced by other men and accepted, and that the "strays" which bother you most are the "strays" from the last reflection from the upper layer. This assumption is just as good as any other. If it does not suit you, find out some other assumption that will suit you better. But whatever the explanation may be, the fact remains that the waves coming from the grinders act like vertically propagated waves, because they affect every part of the receiving system in the same phase and, of course, the signaling waves are horizontally propagated waves.

Mr. Weagant arranges his balance in accordance with this assumption and he proves in a very successful way that he can receive messages practically continuously. Of course, to do that, you have to use long antennas. I felt a little bit unhappy when antennas several thousand feet in length were mentioned. I said to myself "This investigation is evidently being made for a big corporation—no college professor could indulge in anything like that." Then once or twice a generator of two hundred kilowatts, the Alexanderson generator of two hundred kilowatts, was mentioned. Of course, such a machine a poor college professor can never have or possess, and if he did possess it, he would probably sell it and retire for the rest of his life instead of bothering with investigations with such a machine.

The static effects are probably cosmic effects; that is to say, they are terrestrial effects of almost cosmic dimensions. The investigation requires to start with large experimental apparatus,

large facilities for the purpose of producing certain definite effects.

Of course, one idea about the static problem, and I must confess that it is my idea too, is this. If you use a device of the ordinary dimensions that you can conveniently place in any little room, and reduce the static by some means or another, enabling the sender to use about one-one hundredth part of the energy he is usually using, then instead of using anywhere from one hundred to two hundred kilowatts, which almost makes me shudder, he can use from one to two kilowatts. That is my opinion, altho I do not say that my opinion is correct, but in my opinion the real static problem is to reduce the effect of the static by using receiving circuits of very ordinary dimensions, such as we use on board of a ship, and enable one to receive signals free from static interference even if the power of the sending station is reduced one-hundred-to-one. Of course, that is a very, very ambitious proposition.

The proposition described by Mr. Weagant is not so ambitious—it is very much less ambitious. From many points of view it is much more sensible, because advances in an art, as a rule, are made step by step—there is no sudden jump in the development of an art, but a gradual development. Now, this improvement is in the direction of gradual advance. We have here a marked improvement as the first step in the direction of getting rid of the natural interferences of the static, and as such I hail it with delight. It is an accomplishment, a decided accomplishment.

I am surprised, however, that in all this work the name of the vacuum tube amplifier has not been mentioned at all, until towards the last, when the short rectilinear antenna which was described, in which case the energy which was received was very small, and then a five-step to eight-step amplifier was used, which was said to have been designed by the Research Department of the Marconi Company. I would like to see that eight-step amplifier. I have a nine-step amplifier. I know what a tremendous trouble it is to a man to develop a multi-step amplifier, and all of a sudden to hear that some man, who did not say anything about it, had developed an eight-step amplifier. Well, I wonder how many times it amplifies. Ten times, a thousand times, a million times, or what is it?

However, I am glad that Mr. Weagant was finally forced to use the amplifier. I know from private information that as long as you use a long antenna the amplifier is not necessary, because the antenna itself picks up a sufficient amount of energy,

being so long, being such a large trap for the electrical wave to drop in, that an amplifier is not necessary. As long as Mr. Weagant has started to use it, I am satisfied and glad of it, because I do believe that the amplifier is one of the finest inventions in radio telegraphy that we have to-day, and that no problem in radio telegraphy including the elimination of the static can, in my opinion, ever be solved without the full use and full knowledge of the amplifier; and, delighted as I am with the accomplishment recorded in his paper, I am equally delighted with the fact that Mr. Weagant and the Research Department of the Marconi Company have started to use the amplifier in connection with this scheme.

Alfred N. Goldsmith: Several years ago, I had the opportunity of witnessing a demonstration of Mr. Weagant's system of stray elimination under conditions that were entirely under my control. That is, the wiring and connections of the apparatus were completely open; it was permitted to make full wiring diagrams of the equipment, and (after proper instruction) to handle and adjust the apparatus myself. This was the first of a number of opportunities of this sort, and always with the same successful result. It has also been my privilege to have been associated with Mr. Weagant in the development of many phases of the remarkable system which he has originated and particularly in its more recent and advanced forms. In every case, the extreme orderliness of the phenomena presented has been striking and the extent to which theoretical deductions have been verified by experimental evidence was highly gratifying. Particularly has this been the case since, prior to Mr. Weagant's work, strays were regarded as of such random and erratic character that any systematic or logical manipulation of them or their effects seemed hopeless. This state of affairs has been completely reversed, and a powerful weapon of research placed in the hands of pure scientists as well as an instrument of tremendous importance at the disposal of the radio engineers.

Strays are practically omnipresent in radio receivers. Indeed their ubiquitous character has led workers in this field to associate strays and signals as inseparable twins. All the greater is the amazement of the radio manipulator when, on turning a handle, he hears barely audible signals previously smothered by overwhelming crases of strays, emerge finally, loud and well-defined, while the strays dwindle to negligible proportions. The experience, particularly to skilled workers in the art, has

an air of unreality because it is so far removed from all that has previously seemed possible.

With Professor Pupin's view as to the importance of the vacuum tube amplifier *in the development* of Mr. Weagant's methods of stray elimination, I cannot concur. The basic ideas of these methods were quite independent of amplification of this sort, and their main development equally so. Mr. Weagant assuredly owes no debt whatever to the amplifier for his original discoveries and their main reduction to practice.

With Professor Pupin's opinion that the amplifier, in some form, will be used in the more compact stray eliminating systems which Mr. Weagant has more recently developed, I find myself in accord. Convenience dictates receiving antennas of small dimensions. These, being unable to gather considerable energy for the incoming signal, call for amplifiers. As to the amplifiers actually used in this work, the voltage amplification of which they are capable is of the order of Professor Pupin's largest figure rather than of his smallest.

Of the improvements which will result in the radio art because of this advance, little need be said. High speed long range communication and radio telephony assume an entirely new order of importance and a tremendous growth in the radio field becomes the inevitable result of the elimination of the worst obstacle to reliable long distance reception.

No doubt, the auditors of Mr. Weagant's paper have felt the complexity of the methods used, to some extent, particularly at a first hearing. The general impression must be that an ingeniously elaborate electrical means for utilizing the new law which has been found to govern the action of most of the strays which interfere with the reeeption, namely, "grinders," is employed. However, the salient feature of the methods used to eliminate grinders is based on the simultaneity and equality of effects produced by strays on two similar systems properly oriented and in the same horizontal plane. This simultaneous action is the key-note to the situation and, however it may arise, has been the furnace in which the powerful weapon against strays has been forged. The treatment of strays of the "click" class, which are, incidentally, much less serious so far as interference with reception is concerned, has been based on a combination of directional and electrical features ingeniously adapted to the desired end.

Apropos of the differences in the methods of elimination of the two types of strays, one of which is highly vital and the other

glad that it has been my privilege to assist in these pioneer tests.

A little over a year ago, I was ordered by the Department to witness and report on a system of static elimination on test at the Marconi Belmar station. Of so-called static eliminators there had been many in the past, all failing in their purpose, and so my expectations of this new claimant in the field were not great.

On arriving at the station, which was a rough little shack of the standard Marconi coast station type, I saw a number of pieces of apparatus, very crudely wired, and many of obviously home-made origin. It was a cloudy day, with frequent lightning flashes around the horizon, a typical day for "summer static." Two large rectangles of wire supported by the tall masts of the Belmar main station formed two loop antennas, each of these leading into the shack by connecting pole lines. Listening in on one of these loops alone, static was deafening, so loud, indeed, that not a trace of signal of any sort could be heard. Mr. Weagant threw a switch, and even with this crude, preliminary apparatus static died down to a weak murmur, and the signal became clearly readable. I consider that this demonstration was the most impressive one I have ever witnessed.

During the rest of the week, static continued very strong, and many experiments were made. Small concentrated inductances, ground aerials, different forms of tuner connections, and so on, were tried, and in all tests a remarkable increase in signal-static ratio was obtained, whether this ratio was measured in terms of audibility or of readability.

Later tests were carried on at the main Belmar station, in an attempt to make the system more practical for the operator. The first changes lessened the efficiency of the system to a marked degree, but Mr. Weagant persisted in his experiments, trying one thing after another, until the reason for the failure was clearly shown. During this work, I had more than once to fall back on my faith based on the first week's showing.

It became difficult to make further tests at Belmar, as the Navy Department wished to use it for war-time communication, so the experimental work shifted to Miami, Florida. Here Mr. Weagant demonstrated that his analysis of the failures at the Belmar station was correct, and the new Miami circuits worked much better than the original one. As a result of these successful tests, a commercial form of receiving station was erected at Lakewood, New Jersey, the behavior of which during the summer months of 1918 amply justified the claims of the inventor.

Mr. Weagant has omitted the human interest from his account of the Miami work. Yet I can well remember the hours I spent in a drygoods box, miles out in the desert, waiting for the Weagantian command to vary coil, or condenser, or reverse the ever-inverted switch. Again, Mr. Weagant in his paper has referred, quite casually, to the tinfoil-coated pasteboard tubes which covered the connecting leads. Yet the actual construction of this was far from being a casual affair. To direct the activities of fifty colored workers, and to keep them all engaged in placing sheets of tinfoil, one foot square, over six miles of pasteboard tubes, was a task not unworthy of the most determined investigator. At night, especially, this long line of silver, gleaming in the Florida moonlight, seemed more like a monument to Mr. Weagant's persistency than a part of a radio system.

I wish to take this opportunity of referring to the painstaking logical way in which Mr. Weagant has worked out his problem. His skill as an investigator and experimenter has equalled his ability as an inventor. But, above all, he is one of the few investigators who, when confronted with facts at variance with

theory, would dismiss the theory rather than the facts. Such investigators are rare; hence such results are rare.

Ernst F. W. Alexanderson: I feel rather bewildered to discuss such a complicated and deep subject as Mr. Weagant has presented in his paper. Many times when I have been talking with Mr. Weagant about this subject, without knowing the details of his work, I have said: "I certainly do hope that you are right and that the developments will prove to be all that you expect them to be."

Almost every radio engineer must plead guilty of having at some time or other thought that he had a solution for the static problem, only to find later that he was more or less mistaken, usually more so.

Now, what is encouraging, in a very great degree, in Mr. Weagant's paper, is that he gives us a key to the solution by announcing that he has found a new law of nature. Up to the time that I had heard it rumored that Mr. Weagant had found a new law of nature, I was afraid that he might be in a class with the rest who had failed in their efforts. Well, I had a sinking of the heart when he said that perhaps his theory is not altogether proven, perhaps it is not; but the evidence that he has presented to us is so convincing that I hope we will find that a very material advance has been made.

Mr. Weagant, and particularly now Mr. Sarnoff, have pointed out the great importance of a static eliminator in radio telephony. Radio telegraphy has been very reliable during the recent years in trans-Atlantic communication.

I may mention, in this connection, that I happened to be in the New Brunswick radio station, when a call suddenly came from Washington, that a set was needed immediately, and the station operator immediately came in and said that the station was calling Germany. It was the first time since the war that Germany had been called by a United States radio station; and the message that went was the important announcement of President Wilson, stating that the United States could not deal with Germany under its present form of Government, upon which announcement the abdication of the Kaiser followed.

This bears out, further, what Mr. Sarnoff has touched upon, that radio has broken the precedents of international practice, by permitting direct communication between the responsible parties in the belligerent nations, thereby short-circuiting the usual channels of diplomacy.

President Wilson is now on the sea, and arrangements have been made whereby telephone messages are being sent to the President every day. This apparently has nothing to do with Mr. Weagant's paper directly, but Mr. Sarnoff's discussion leads the thought from one to the other, and that is that the use of radio telephony across the ocean is limited in its possibilities by the degree to which static can be eliminated.

I wish I had with me a photograph which I was examining on the train; a photograph of the radiated waves from the New Brunswick station—being the electrical equivalent of Secretary Daniel's voice when he was speaking to President Wilson on his way over. We hope that in the future, when the best kind of receiving devices are in existence on both continents, that many such photographs will be taken of the rulers of the world.

David Sarnoff: Some time ago, I asked Mr. Weagant to tell me, if he could, the particular thought or idea responsible for his faith in the ultimate solution of the static problem. I asked the question specifically, because of the apparent disbelief of so many others that a real solution of this vexatious problem could be obtained.

In answer to my question, Mr. Weagant stated that he had always considered Nature reasonable and logical; it followed, therefore, that it would not, on the one hand, bestow upon mankind a boon, such as electrical communication thru space; and, on the other hand, place in its way a deadly barrier such as static has been, without offering means of nullifying it and attaining the full advantages that space communication offers to the world.

It was this implicit faith in the justice of Nature which spurred Mr. Weagant on in his determination to master the disturbing elements. The task, has, perhaps helped to add a few gray hairs to his otherwise young head. He has told you himself how he reached his goal, and I merely wish to call attention to the original inspiration and conviction, characteristic of the man.

In my judgment, the elimination of static interference marks the most important practical advance in the radio art since Marconi's original invention.

International radio telegraphic communication, a child of the past, will now grow rapidly to sturdy manhood. Radio telephony over long distance and across the oceans—impracticable heretofore—is now in full view, and commercial radio telephone service between the United States and Europe may confidently be expected.

Think what this means. Electric signaling, now more than three-score years old, has not provided means for talking to our friends across the great oceans. Whatever we had to say, others said for us by telegraph code. And now, for the first time in the history of electrical science, the spoken word may be uttered by us in our own language and heard by the desired ears across the oceans. I predict that trans-oceanic radio telephony will in time revolutionize international business, and diplomatic and social intercourse in the same way that the Bell telephone revolutionized our daily affairs on this continent.

Mr. Weagant made reference in his paper to the possibility of conducting long distance radio communication with less power at the transmitter than is now generally employed. This, it seems to me, should logically follow as one of the results of his great invention, and one is now justified in expecting that, before long, communication across the Atlantic may be carried on successfully with transmitters of, say roughly, fifty kilowatts, or perhaps less, and receivers of the compact type described by Mr. Weagant.

Nothing brings nations and peoples closer together than reliable, rapid, and cheap communication, and radio now promises to be the international courier, fulfilling these three vital requirements.

The present high cable rates between widely separated countries have limited the amount of news or press matter exchanged between the United States and such countries as, for example, China, Japan and Australia. The mail service is, of course, too slow to record important events.

With the elimination of static interference and the possibility of reduced power at the transmitters, it is conceivable to me, and no doubt to many others, that two or three long distance transmitting stations, located in the most important and suitable parts of the world, could be devoted to the exclusive transmission of daily news or press matter, broadcasted to all the countries, where, with the use of the proper receiving system, the broadcast messages could be received by all and published in the press of the world.

Cable companies and the interests they represent have long made use of their favorite argument that communication by radio is not secret, and whereas by cables it is. Of course, I need not tell you practical men that no system of communication is really secret; but the very fact that several transmitting stations can simultaneously communicate with the entire world, gives to

radio an advantage that the cables never had and never can possess.

Philip E. Edelman (by letter): Having recently completed similar work with different means I was naturally much interested in Mr. Weagant's paper, and would like to ask Mr. Weagant whether he has been able to balance out strays other than "grinders," and if "clicks" are eliminated or not?

As described, apparently the plan would be limited to a fixed wave length to keep the correct loop spacing, but doubtless could be worked out into the same flexibility as the usual old style receiving stations.

The correct explanation seems to me to be that the single turn loops shown are directional with respect to strays as well as signals. Experiment shows that a vertical loop oriented east and west does not usually receive the same strays as one in the same position, but oriented north and south, does. If strays came only directly from above and below, as seemingly stated, it would appear that two loops placed at right angles to each other would also get the same strays simultaneously. Experiment shows that this is not the case, for simultaneous records prove that one loops receives strays the other does not. Accordingly, it makes no difference where the strays come from, because all three loops are oriented the same and receive only such portion of the strays as come within their directional locus. All strays originating from a sufficient distance can accordingly be balanced out whether they come from overhead or under foot or at any angle in the receiving cone of the loops.

The loop apparently owes its directional property to the fact that for maximum induced current therein, it is essential that the advancing waves cut the turn of the loop at right angles thereto, which means that the axis must be in alignment with the shortest distance to the transmitting station or source of strays. Waves from other sources at other angles to the axis have a lesser effect, roughly (*Maximum*) $\cos a$, where a is the angle, until at right angles there is no effect and this proves to be the case experimentally. Accordingly part of the stray mitigation is due to the fact that strays arriving from sources outside of the cone of the loop have slight if any effect thereon in the first place. Such as do affect the three loops in alignment apparently affect all alike and can be balanced out. A similar argument would appear to hold for a linear ungrounded antenna such as Mr. Weagant shows.

I would like to ask what operating ratios of signal-to-stray audibilities were obtained under the new method of measurement outlined and how they compared with the usual old method of audibility measurement. The latter is notoriously not precise because depending upon the sensibility of the operator's ear which varies widely in different people.

This is an excellent demonstration of stray mitigation, but still leaves many problems, not the least of which is the so called "fading" effect. Even if total stray elimination could be effected at the receiver, the media between the transmitter and receiver remain outside control, so that signals can still fade erratically due to fluctuations therein.

A NEW METHOD OF USING CONTACT DETECTORS IN RADIO MEASUREMENTS*

By

Louis W. Austin

(United States Naval Radio Laboratory, Washington)

For many measurements in radiotelegraphy it is necessary to use a radio frequency current indicator of known resistance. If the current to be measured is small, it is generally customary to use a thermoelement and galvanometer. The most sensitive thermoelements are either of the vacuum type or the welded tellurium type. The vacuum thermoelements can be obtained of any desired resistance and are very sensitive, but are slow in action and frequently show a bad zero drift. In addition, the deflection usually shows considerable divergence from the current-square law. The tellurium platinum elements are quick acting and follow the current-square deflection law with sufficient accuracy for all practical purposes. They are, however, so fragile and difficult to manufacture and transport that no manufacturer has yet undertaken to supply them commercially. It is also impossible to make the contact resistance much less than 10 ohms. It is to be noted that the resistance in both the vacuum and tellurium types changes considerably with the amount of current flowing.

On account of the difficulties mentioned, the sensitive thermoelements in our laboratory have been replaced, for the most part, by a shunted contact detector circuit arranged as shown in Figure 1. Here $L\,C$ is any oscillating circuit having inductance and capacity, D is a contact detector, G a high resistance galvanometer, K a paper condenser of one microfarad capacity, and R a resistance which may have any value from 0.1 to 100 ohms. The greater part of the radio frequency current passes thru R, while a small portion is shunted thru the condenser K and the detector. The direct current from the detector after passing thru the galvanometer returns thru R. On account of the high resistance of the detector, the total resistance of the detecting

* Received by the Editor, December 23, 1918.

system is practically identical with R, as has been experimentally tested between 0.1 and 100 ohms.

The sensibility of this arrangement is much greater than that of the best vacuum thermoelements of equivalent resistance.

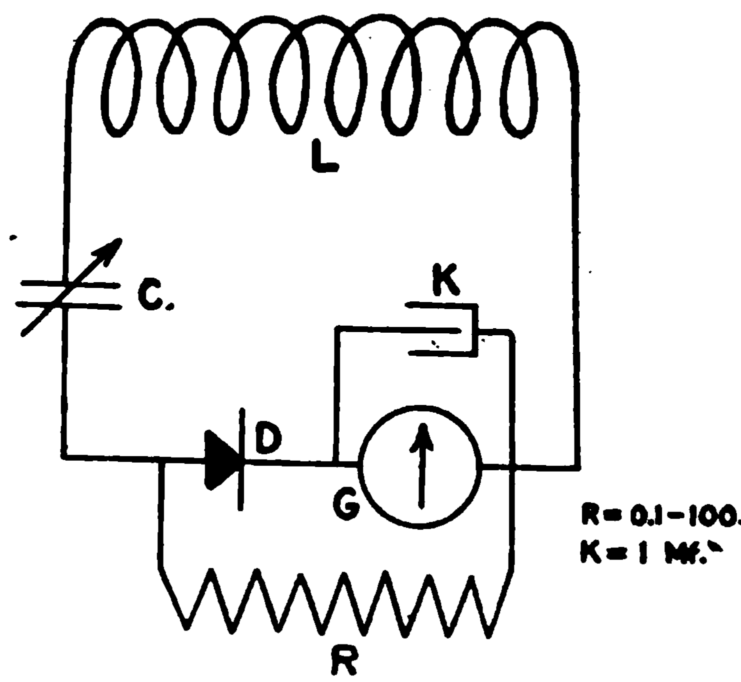

FIGURE 1

In the case of most of the well-known detectors the proportionality between deflection and current-square is excellent. Galena, while the most sensitive of any of the detectors tried, shows a slight deviation from the square law. For absolute current measurements the system may, of course, be calibrated by comparison with a known thermoelement at the time of experiment.

Since the fraction of the total current passing thru the detector is, for a wide range, practically proportional to the shunt resistance, it is possible to calibrate the apparatus approximately with a 1-ohm shunt which, with a galvanometer of a sensitivity of 5×10^{-9} amperes, and with an average silicon detector gives a deflection of 1 millimeter for about 2×10^{-3} amperes radio frequency current. Thus 300 millimeters on the galvanometer scale represent about 34×10^{-3} amperes. This can be read conveniently on the small hot-wire instruments found in most laboratories, which give full scale deflection for from 80 to 100 milliamperes. The shunted detector can then be used for other shunt values by dividing the sensibility by the shunt ratio.

The following table gives the approximate radio frequency current required for 1 millimeter deflection on a galvanometer having a direct current sensibility of 5×10^{-9} amperes. If the sensibility is 5×10^{-10}, 1 millimeter with a 100-ohm shunt will

represent approximately 6×10^{-6} amperes, radio frequency current. The third column gives the maximum radio frequency current which can be safely sent thru the system without danger of materially changing the detector resistance. The values are based on a detector resistance of 3,000 ohms and a maximum safe detector current of 12×10^{-6} amperes.[1]

TABLE I

Approximate R. F. Sensibility and Maximum R. F. Current for Various Shunts with a Silicon Detector.

Shunt (Ohms)	R. F. Sensibility (10^{-6} Amperes)	Maximum R. F. Current (10^{-6} Amperes)
1	2,000	36,000
5	400	7,200
10	200	3,600
25	80	1,400
50	41	710
100	21	360

Of course, much larger currents can be used with good proportionality between deflection and current square provided exact constancy of resistance is not required.

In using this circuit care may be taken that there is no direct action of the outside driving circuit on the detector shunt loop *D K R*.

SUMMARY: An arrangement for using crystal detectors in radio measurements is shown. It is based on the original use of a low resistance shunt across detector and galvanometer and calibration of the arrangement using r.f. currents which can be measured with hot-wire instruments. By increasing the shunt resistance, the necessary much higher sensibility is directly obtained. Performance data are given.

[1] See "Contact Rectifiers of Electric Currents," "Bulletin—Bureau of Standards," volume 5, 1908, page 133, Reprint 94.

THE POSSIBILITIES OF CONCEALED RECEIVING SYSTEMS*

By

A. Hoyt Taylor

(Professor of Physics, University of North Dakota)†

No one who has had any considerable experience with continuous wave receivers can fail to note the remarkably loose coupling which may be successfully employed between the antenna circuit and its secondary.

Some time ago, it occurred to the writer that even for long distance reception it ought to be possible to dispense with the antenna and ground connection by expanding the secondary circuit into a form which would cut sufficient magnetic lines in the wave to give, with a sensitive receiver, readable signals. The excuse for reporting these experiments is that while the principles involved are not new, the results obtained have been rather surprising, and indicate possibilities which may have been overlooked by some earlier experimenters.[1], [2].

The secondary circuit of the receiving set at "9XN" (Grand Forks, North Dakota) was accordingly replaced by a rectangle 10 feet (3 meters) square, of 40 turns of number 27 double cotton covered wire.[3] The rectangle was hung up inside a room in a brick and steel building which is full of wiring conduits and gas, water, and steam piping.

Audible signals from "NAA" (Arlington, Virginia), "WSL" (Sayville, Long Island), "WGG" (Tuckerton, New Jersey), and "NAJ" (Great Lakes, Illinois), were received.[4] The turns

* Received by the Editor, July 8, 1916. This paper will be followed in early issues of the Proceedings by three papers by Commander Taylor on the use of ground wire systems. the elimination of strays, and remote control stations.—Editor.

† Now Lieutenant-Commander, United States Navy.

[1] F. Braun, "Jahrbuch der drahtlosen Telegraphie und Telephonie," January, 1914.

[2] Pickard, "Electrical Review," 50.

[3] Diameter of number 27 wire = 0.0142 inch = 0.36 mm.

[4] (The distances from Grand Forks of each of the stations mentioned are respectively 1,210 miles (1,950 km.), 1,190 miles (1,915 km.), 1,170 miles (1,885 km.), and 620 miles (995 km.) practically all over flat country, one moderately high mountain range intervening in the first three cases. Arlington is equipped with a 60-kilowatt arc set and a 100-kilowatt spark set, Sayville with a 100-kilowatt alternator-frequency changer set, and Tuckerton with a 60-kilowatt arc set.—Editor.)

of this rectangle were evidently too close together, giving a bad distributed capacity effect. A new rectangle with the wires wound side by side in a flat band gave better results, and a third one with 16 turns spaced about 2 millimeters (0.08 inch) apart gave such satisfactory reception of all waves from 2,500 meters up that it was decided to continue the experiments at the writer's home where the rectangle was hung up in a tree and arranged so that it could be rotated so as to take advantage of the directive effect. It was found unnecessary to have the lower wires more than 5 feet (1.5 meters) from the ground.

Figure 1 shows the complete receiving system. The adjustments are very simply made by simultaneous variation of the condenser C and the plate circuit inductance L. The bulbs used

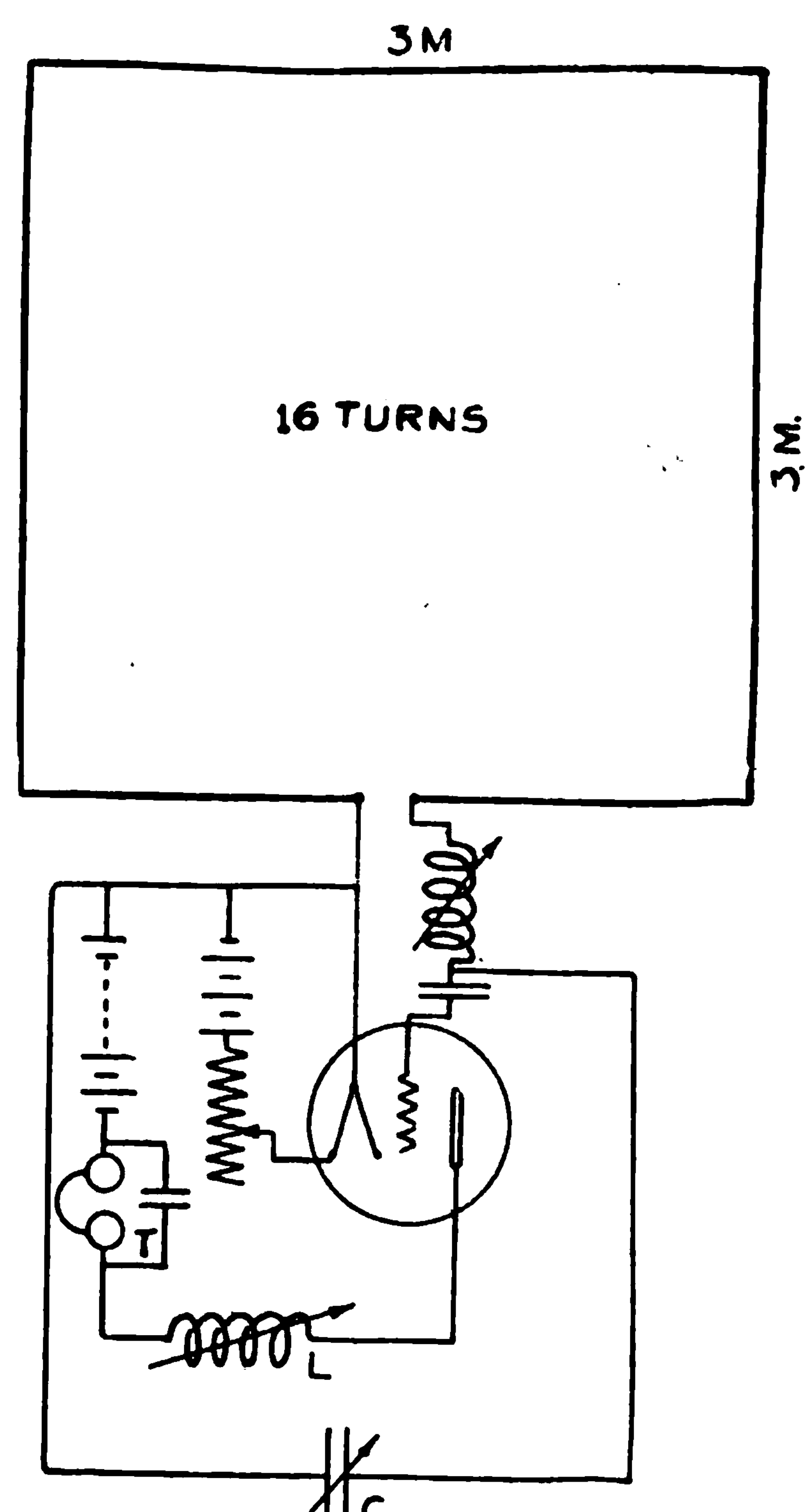

FIGURE 1—WIRING OF RECEIVER

were cylindrical, with cylindrical plate, spiral grid, and two straight filaments, only one of which was used. Nineteen of these bulbs were tested, and only one found which with proper

choice of high potential battery and heating current failed to give satisfactory results. The voltages used on different bulbs were found to vary between 22 and 40 volts, and the filament current from 0.78 ampere to 0.90 ampere. For any one bulb, the adjustments were exceedingly constant and certain of duplication. The writer regrets to report that attempts to measure audibilities by the shunted telephone method, using a Brandes 3,200 ohm telephone receiver, were not very successful, since the bulbs began to give the siren effect as soon as the shunt was reduced to between 100 and 300 ohms. Shunts were occasionally made with two receivers in series. For Arlington working on 6,000 meters by daylight, this shunt was usually about 200 ohms when they were working with "NBA" (Darien, Panama Canal Zone). The writer is of the opinion that such shunts cannot readily be translated into audibilities, since shunting a receiver unquestionably affects the conditions of oscillation in the bulb, and the effectiveness of the shunt depends on the pitch of the signal observed. Bulbs giving equally loud signals will shunt down to very different values.

Observations were begun in the latter part of April, 1916, and daylight signals from Arlington, Virginia; Key West, Florida; Darien, Panama Canal Zone; Sayville, Long Island; Tuckerton, New Jersey; Great Lakes, Illinois; New Orleans, Louisiana; San Diego, California; Bolinas, California, and South San Francisco, California, were always readily readable when the plane of the rectangle coincided with the direction from which the waves arrived. The distance from Darien is 3,000 miles (4,800 km.). During June and July, the strays prevented the reception of Bolinas and very often of South San Francisco. The most remarkable performance was the almost daily reception of the 2,500 meter mid-day time signals from Arlington up to date (July 6). The tone of the spark was destroyed, as it was absolutely necessary to have the bulb oscillating in order to get the signals.

The 7:30 P. M. Central time press reports sent out on the same wave length were copied nearly every day up to June 15, altho it is broad daylight here at that hour.

In connection with these results it may be pointed out that the efficiency of a receiving system in the period of summer strays depends not so much on the absolute audibility of the signals as on their relative audibility as compared with strays and other disturbances. In this respect, the rectangle has a very great advantage. If the strays are equally distributed

from all points of the compass, the rectangle, owing to its directivity, cuts out a considerable part of them.

The selectivity may of course be greatly increased by using a variable condenser in series with the rectangle and adding an inductive coupler. The set then becomes a standard set with the antenna and ground replaced by rectangle and variable condenser. The signals are about 50 per cent. stronger when all adjustments are carefully made, and the strays are a little weaker. Such an arrangement was used when working at night, when an arc light less than 300 feet (100 m.) distant created bad interference below 6,000 meters. With this standard and well-known arrangement, however, the set loses the prime advantage of simplicity. The operation involves just twice as many adjustments.

Experiments were continued with two other rectangles, each of sixteen turns wound edgewise, or in the plane of the rectangle. One rectangle, having an area of 103 square feet (9.57 sq. m.) was hung against the east and west wall ot a second floor room, and the other, of 77 square feet (7.15 sq. m.) area was hung against the north and south wall. Figure 2 shows the connection used. By closing the switch S_1 and throwing the switch S_2 to the left, the east-and-west rectangle could be used; by closing S_3 opening S_1 and throwing S_2 to the right the north-and-south rectangle was in circuit; by opening S_1, throwing S_3 to the right and closing S_2 to the left both rectangles were in service, and so connected as to have directivity to the southeast. By reversing the connections of the north and south rectangle (by throwing S_3 to the left) the resultant directivity could be changed from northwest and southeast to southwest and northeast. With this arrangement signals from all points of the compass could be received with a fair degree of directivity.

This set permits most of the selective advantages of the single rotary rectangle in a convenient and easily concealed permanent installation. The rectangles may be readily concealed behind a tapestry or, if need be, inside of the wall, except in the case of steel structures where they could not be used inside for reception over distances greater than 500 miles (800 km.) depending on the type of building and its surroundings.

An example of the usefulness of the two rectangle combinations may be cited. When Darien 19 degrees southeast from Grand Forks is sending at the same time as south San Francisco 58 degrees southwest, there is no interference even on the same wave length if the switch S_3 is thrown so as to give southwest

directivity in the one case and southeast in the other. Similar results were obtained in eliminating interference between Arlington and Bolinas. The southeast combination eliminates San Diego while the southwest combination greatly weakened Arlington signals altho they were still readable. If Arlington and San Diego were sending simultaneously, they could readily be separated.

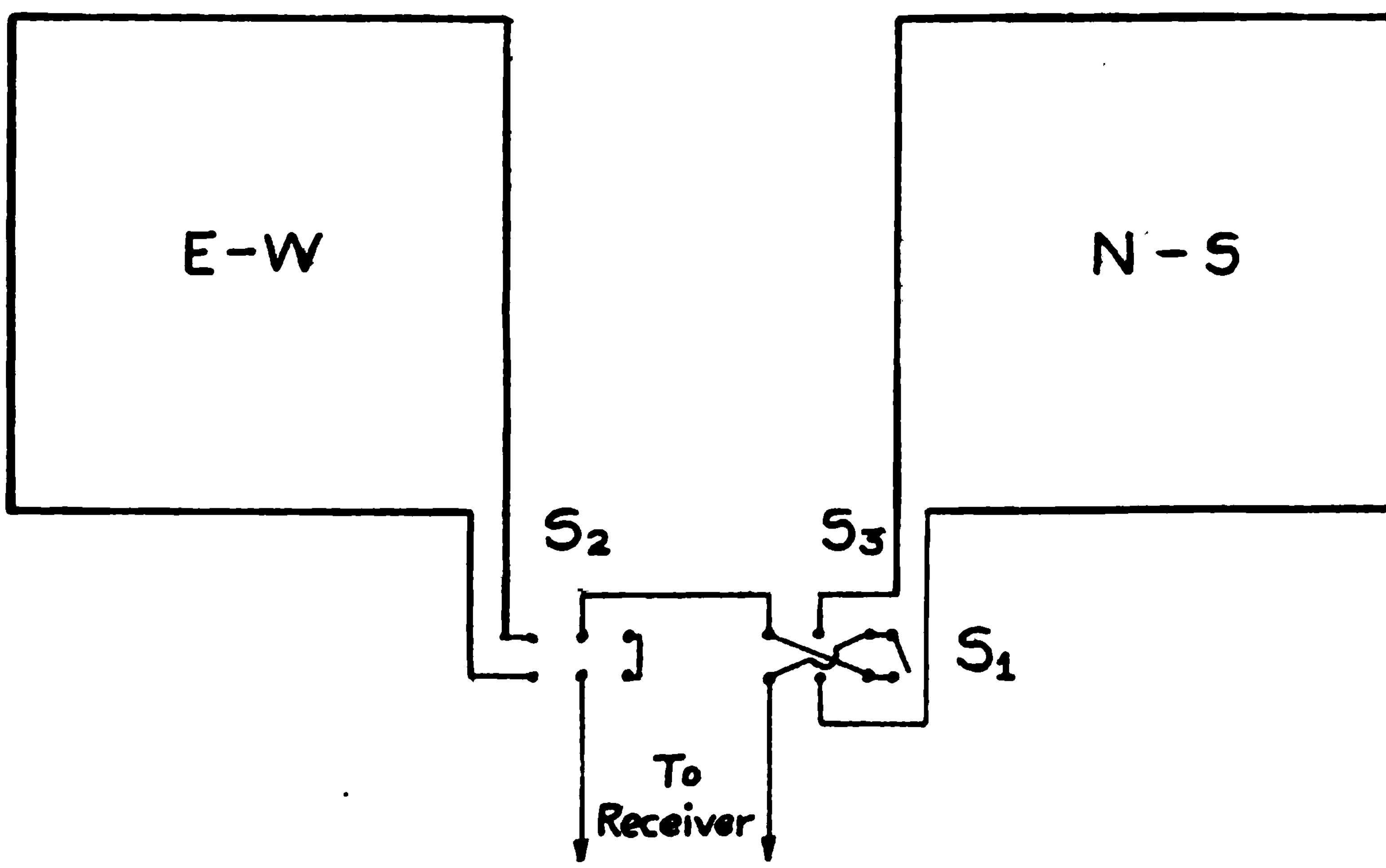

FIGURE 2—DOUBLE LOOP ARRANGEMENT

For best results, the two rectangles should be separated so as to cut no common flux, altho this was not done in the experiments here described.

With a single rectangle of 16 turns, a loading coil was used above 3,500 meters, and also with two rectangles above 5,000 meters. A larger and variable number of turns on the rectangle will permit the omission of the loading inductance and a maximum reception at all waves. The latter is of no great advantage after a certain number of turns have been reached, as the gain in signal strength is no longer proportional to the number of turns. Sayville, distant 1,170 miles (1,885 km.), could be read with 5 turns, on one rectangle, in the evening, during April.

Reception with the rectangles was compared from time to time with reception on the regular set at Grand Forks, using the 800-foot (243 meters) three-wire antenna the far end of which is 120 feet (37 meters) high, and near end 60 feet (18 meters) high. Altho the signals at Grand Forks are very much louder,

they are, with few exceptions, not so readable thru the summer strays.

The advantage of the set here described may be summarized as follows; it is cheap; compact; immune from storm damage; sufficiently sensitive; thoroly reliable; simple in adjustment; and, by its directivity, partially eliminates strays and reduces interference. Moreover, it is readily concealed, if, for military or other reasons, it is desirable to do so.

SUMMARY: A closed 3-meter square loop of 16 turns of wire, hung about 1.5 meters from the ground, was used for long distance reception with an oscillating audion. Daylight overland reception from stations as distant as 3,000 miles (4,800 kilometers) was regularly accomplished.

A combination of two similar loops was also employed, and considerably increased the directional selectivity.

The inherent directional qualities of these receivers were utilized in the reduction of strays and interference.

ON MEASUREMENT OF SIGNAL STRENGTH*

By

W. H. Eccles, D.Sc.

(Professor of Applied Physics and Electrical Engineering, City and Guilds of London Technical College, Finsbury, England)

The theoretical discussion below has been written with the object of emphasizing the difficulties that arise in the interpretation of the shunted telephone method. Besides this difficulty of interpretation, it is notorious that the measurements by the shunted telephone are not susceptible of accuracy. Now that the War is over, it is to be hoped that new measurements by improved methods will be attempted and that those to whom will fall the great opportunity of carrying out the work will select methods capable of yielding trustworthy information.

The American investigators of the strength of signals from great distances have made much use of an aural method of measurement called variously the "shunted telephone" or "parallel ohm" method. The measurement consists in connecting a variable resistance across the telephones and finding the value of the resistance that will just reduce the sound of the signals to a standard intensity called the unit audibility. This value substituted in a formula gives a number called the "audibility factor." "Unit audibility" is defined for any given set of apparatus as that at which dots and dashes may just be discriminated. The formula used by L. W. Austin and by J. L. Hogan for calculating the audibility factor A from the value of the shunt is

$$A = \frac{(R+S)}{S} \qquad (1)$$

where S is the resistance of the shunt in ohms and R the impedance, or perhaps the resistance, of the telephones. There is some doubt about accepted usage with regard to the symbol R, but, as will be shown later, the use of impedance or resistance is less material than will at first sight appear. Where it is helpful to distinguish the two cases we may write

* Received by the Editor, February 25, 1919.

$$A' = \frac{(Z+S)}{S}. \qquad (2)$$

The typical circuit for the measurement of the audibility factor is given in Figure 1. It will be obvious that the chief effect of varying S is to change the proportion in which the pulsating current from the detector divides between S and the telephone. But, in addition, varying the shunt alters the resistance in series with the detector, and the shunt, it will be noticed, permanently affords for radio oscillations a path additional to the condenser C_3. However, to remove this defect, it has been proposed to insert a choking coil in the lead between C_3 and the telephone at the point marked χ. Sometimes instead of including a condenser C_3 in the detector circuit, the capacity of the telephone leads and windings is entrusted with the task of passing the radio oscillations, the telephone then replacing C_3.

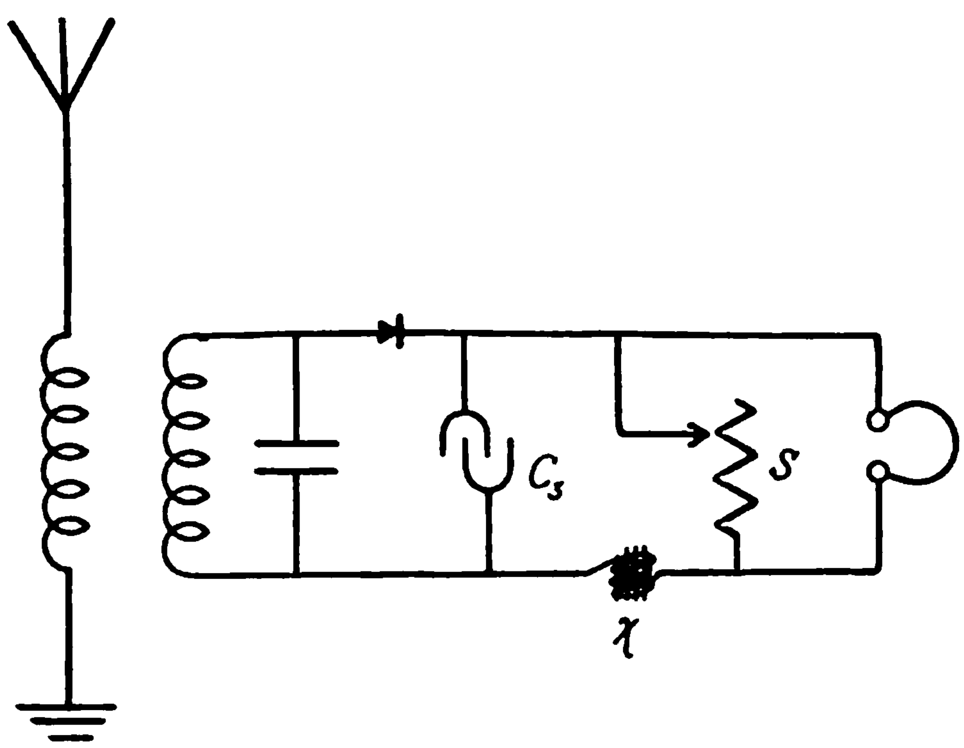

FIGURE 1

Let us imagine that continuous waves are being received and heterodyned and that an audible note is being produced in the telephone. The question arises: How does the audio frequency current distribute itself between the telephone and the shunt? We shall investigate this problem by supposing that an audio sine voltage of amplitude V exists at the terminals of the parallel circuit, superposed upon a steady component which may be disregarded in the analysis.

The circuit and its vector diagram are drawn in Figures 2 and 3, the circuit being regarded for the purpose as an auto-transformer with resistance coupling, the common part being S with a

current $I+J$ traversing it, the telephone having a current J thru it, and the sine current of amplitude I being the current from the detector smoothed by condenser C_3 and choking coil χ.

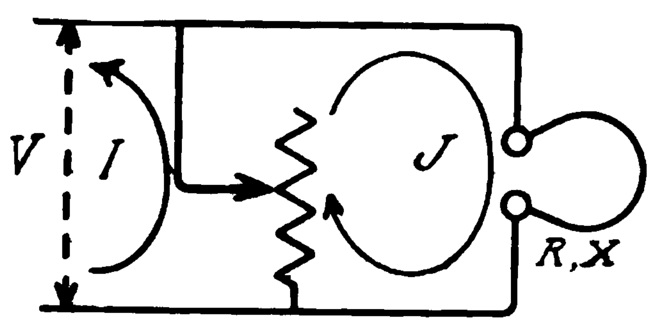

FIGURE 2

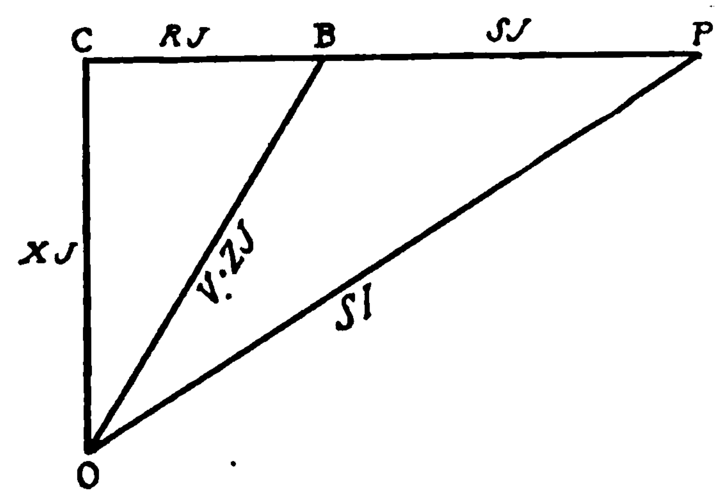

FIGURE 3

First the triangle OPB for the I circuit is drawn; it may be of any convenient shape for trial so long as the vector SJ is shown less than SI. In the triangle the potential drop SI is represented by OP and is shown as made up by the applied voltage $OB=V$ together with the reaction $BP=SJ$ from the secondary circuit. This latter circuit is represented by the right-angled triangle OPC, wherein the potential drops are $PB=SJ$, $BC=RJ$ and $CO=XJ$, where X is the audio reactance of the telephone. These are shown made up by the action $PO=SI$ from the I circuit. From this triangle it is clear that OB represents ZJ as well as V, as is otherwise evident.

The reactance of a telephone has been studied in great detail by many observers, and especially in recent years by A. E. Kennelly, G. W. Pierce, and H. A. Affel. From these researches it is known that the resistance of a telephone increases with frequency, while the reactance undergoes fluctuations near frequencies related to the natural frequencies of the diafram. The value of the reactance is never zero or negative, and as the audibility factor in a definite set of experiments is concerned with only one frequency, we may treat both R and X as positive constants.

Now in the practice of the method the telephone current J is always brought to the same value, namely, that giving "unit audibility." Therefore OC, CB are constant, and as S is varied, BP varies in simple proportion. The ratio of OP to PB is the ratio $\frac{I}{J}$. It is the ratio of the current from the detector thru the parallel circuit to the current giving unit audibility. It

will be called the strength ratio a, so that $a=\frac{I}{J}$. The ratio $\frac{\mathrm{CP}}{\mathrm{PB}}=\frac{(R+S)}{S}=A$, the audibility factor according to equation (1).

When the current I from the detector is just equal to J the strength ratio is unity and S must be infinite. Then P is at infinity and OP is parallel to BP. Also BP and CP are both infinite and the limit of their ratio is unity, that is the audibility factor is unity. At the other extreme, when the current from the detector is very great, S must be a very small resistance and therefore P is very near to B. The ratios $\frac{\mathrm{OP}}{\mathrm{PB}}$ and $\frac{\mathrm{CP}}{\mathrm{PB}}$ are then both infinite. The strength ratio is not equal to the audibility factor except near the limit $A=1$.

The diagram enables the connection between a and A to be expressed easily. It is only necessary to write down the well-known trigonometrical equations for the cosine of the angle OPC in terms first of OP and OC and then in terms of OP, PB, and OB in order to obtain the relation

$$a^2=\left\{\frac{(A-1)Z}{R}\right\}^2+2A-1 \tag{3}$$

The equation is exhibited as a hyperbola in Figure 4, which by its departure from the dotted line indicates that the value of A, the audibility factor, is in general considerably different from that of a, the strength ratio.

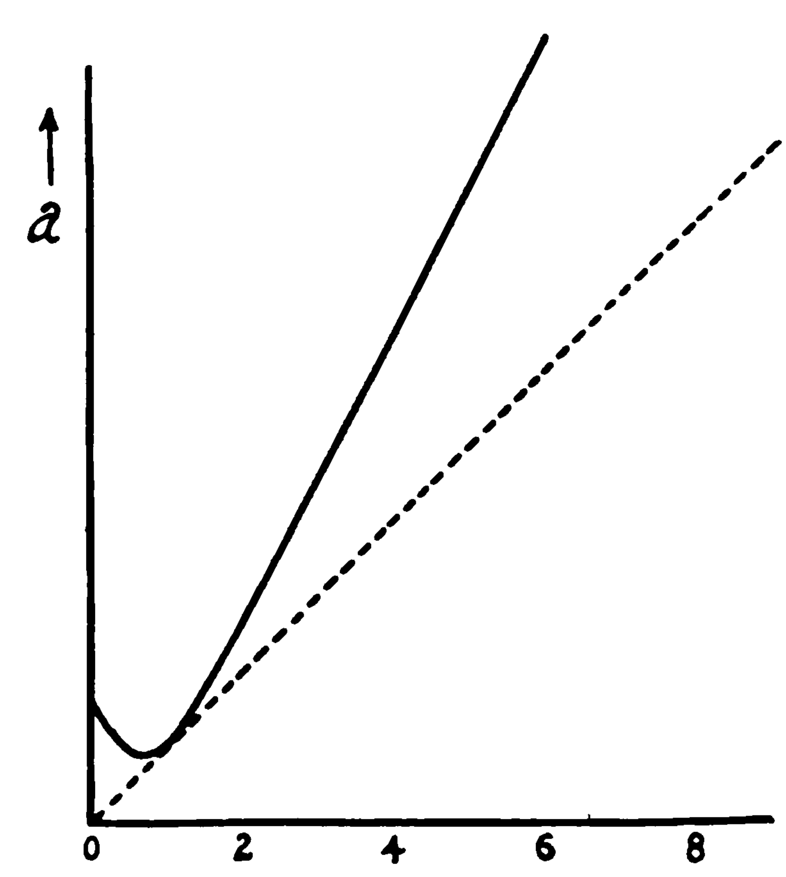

FIGURE 4

Austin, Hogan, and others have therefore used instead of equation (1) as the definition of audibility factor the equation

$$A' = \frac{(Z+S)}{S} \tag{2}$$

In order to examine the connection between them we rewrite the equations in the form

$$S(A-1)=R, \quad S(A'-1)=Z \tag{4}$$

and then by division obtain

$$A'-1=\frac{(A-1)Z}{R} \tag{5}$$

Then by substitution in equation (3) we get

$$a^2=(A'-1)^2+\frac{2(A'-1)R}{Z}+1 \tag{6}$$

This is a rectangular hyperbola and is traced in Figure 5. Comparison with the dotted line proves that the revised definition of the audibility factor still differs considerably from the strength ratio. However, since either audibility factor is readily com-

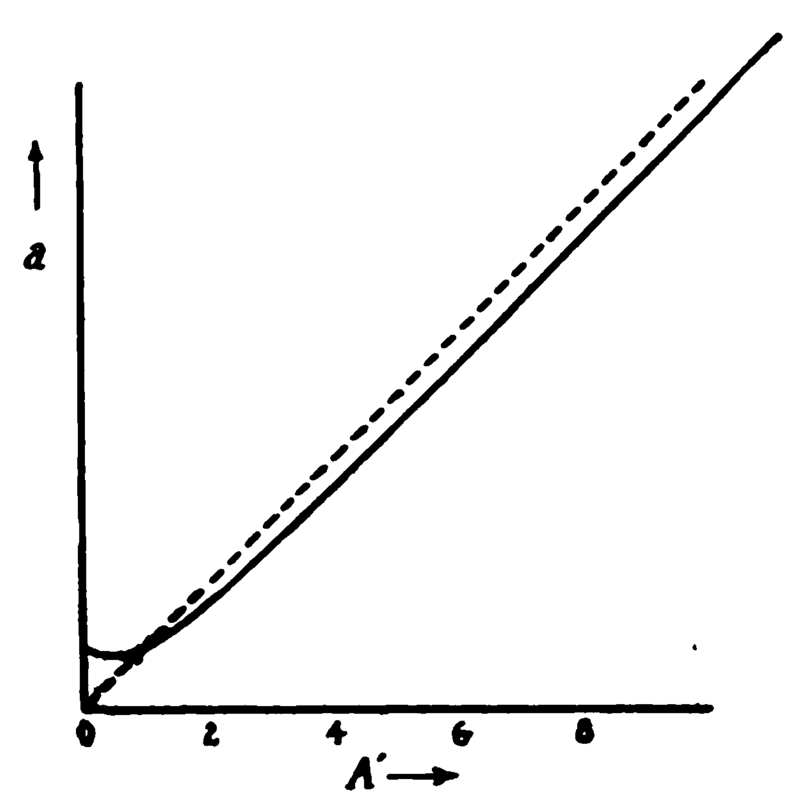

FIGURE 5

puted, it is best, when the strength ratio is required, to compute A or A' first, and then deduce a by aid of the appropriate formula.

This method of measuring audibility can be applied legitimately to the determination of the variation of the strength of received signals from a moving sending station such as a ship, but it is not suitable for the comparison of different fixed sta-

tions having different note frequencies. In what follows, it will be supposed that the application is made to problems of the former kind. It is then necessary to examine what radio quantity the audibility factor and the strength ratio actually measure.

First it must be noticed that since the telephone current J is always adjusted by ear to the same value, the terminal voltage V is always the same. Therefore the total work done in the telephone and shunt, which is $\frac{1}{2} VI \cos\phi$, where ϕ is the phase angle of Figure 3, depends on the first power of I. The power absorbed by the antenna is $\frac{1}{2}(R_1+R_1')I_1^2$, R_1 being the antenna resistance, R_1' being the image in the antenna circuit of the detector resistance, and I_1 being the antenna current amplitude. The proportion of this power passed to the detector is $\frac{1}{2}R_1'I_1^2$. Let η be the efficiency of conversion of radio energy to audio energy by the detector. Then we have by equating the expressions obtained above

$$V I \cos\phi = \eta R_1' I_1^2 \tag{7}$$

The image resistance R_1' depends on the couplings of the circuits between the antenna and the detector, and therefore these ought always to be rigidly constant in the application of the method. Then R_1' and V are the constants in this formula, and introducing a, the strength ratio, we may put $a\,J$ for I. We thus obtain

$$I_1^2 = \frac{V J}{R_1'} \frac{a \cos\phi}{\eta} \tag{8}$$

If η were constant the energy collected by the antenna would be represented perfectly by $a \cos\phi$ and the antenna current would be proportional to the square root of $a \cos\phi$; but η is not constant, tho for loud signals the efficiency is practically constant, and then we conclude that the square of the antenna current is measured by the quantity $a \cos\phi$. This may be evaluated from Figure 3 by expressing the value of the cosine of ϕ, which is the angle BOP, in terms of the lengths of the vectors. We obtain the equation

$$a \cos\phi = \frac{a^2(S^2+Z^2)-S^2}{2aZS} \tag{9}$$

which is easily evaluated numerically from experimental data. For small values of S, that is for loud signals, we may say that approximately

$$a \cos\phi = \frac{aZ}{2S} \tag{10}$$

For large values of S, that is for faint signals, we may write approximately

$$a \cos \phi = \frac{(a^2-1) S}{2aZ}.$$

The quantity $a \cos \phi$ may be called the audio power ratio. If the detector were of the same efficiency for all magnitudes of radio current, this quantity would also be the radio power ratio. But until the efficiency is thoroly investigated at low powers, the extant measurements on the propagation of signals to great distances cannot be confidently interpreted.

Some progress towards this interpretation has been made by B. van der Pol. Damped trains of oscillations were induced in a typical receiving circuit by means of a variable magnetic coupling with a circuit in which constant oscillations were being produced. The calibration of the mutual inductance gave the relative magnitudes of the oscillatory current passed to the detector at various settings. The receiving circuit contained a shunted telephone and measurements of the audibility factors were made at all settings. The results are seen in Figure 6 from van der Pol's paper in the "Philosophical Magazine" of September, 1917. From abscissa 0.6 (where $A=4$) to 2.2 (where $A=160$),

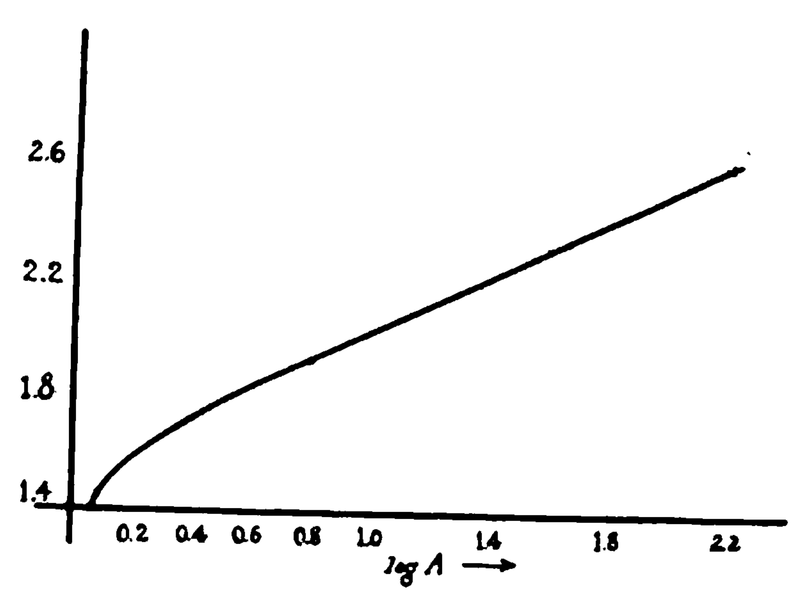

FIGURE 6

the curve is straight with gradient 0.5. This indicates that A is proportional to the square of the radio current delivered to the detector, or that the audibility factor is a fair measure of the signal energy for medium strengths. With weaker signals, however, while A ranges from 4 down to 1.2, the logarithmic curve bends downwards so as to pass thru the origin corresponding to unit audibility. During this stage the audibility factor is proportional to powers of I gradually changing from 2 to 0.7.

It has been pointed out by G. W. O. Howe that the use of A' instead of A, that is to say, the use of the telephone impedance instead of its resistance in the calculations, would bring the straight part of the curve slightly lower on the diagram. The straight parts are, however, not of great interest. The important part of the curve, for the purpose of interpreting the results of measurements on long distance signalling, is the curved part. Since A and A' are connected by the linear relation (5), the substitution of one for the other leaves the general appearance of the curve unchanged tho tending to straighten it a little. The effect of using $a \cos \phi$ is, however, significant. The relation between these modes of measuring is best seen by redrawing Figure 3 so as to exhibit $\mathrm{OD} = \mathrm{OP} \cos \phi$, and, after dividing every line by SJ, marking them with their values. Then keeping the unit line constant and imagining S to decrease a little, we get the consequent increases of A, a, and OD as marked in Figure 7. Clearly $\partial\,(\mathrm{O\,D}) > \partial\, a > \partial\, A$. The longer the unit, that

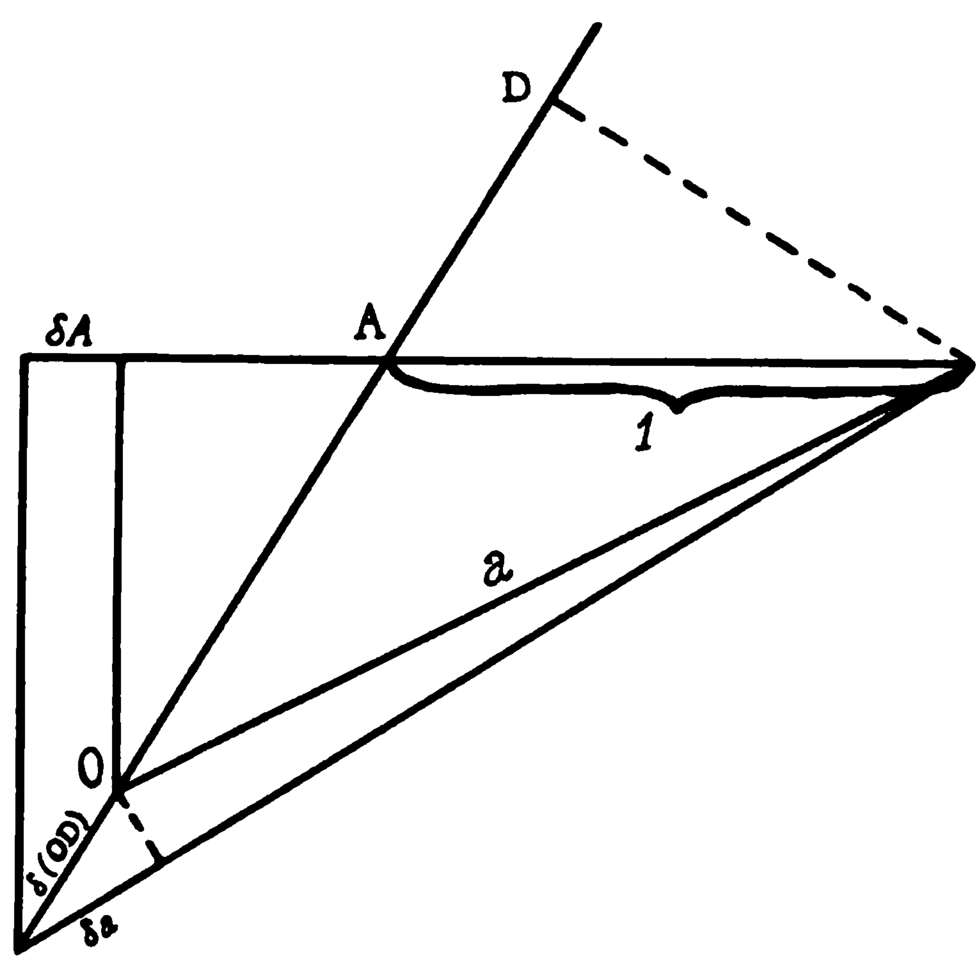

FIGURE 7

is the weaker the signal, the more does ∂ (OD) exceed $\partial\, a$, but the less does $\partial\, a$ exceed $\partial\, A$. This is indicated in Figure 8. The curves a and $a \cos \phi$ tend to approach and run together when signals get very strong.

The main result is that van der Pol's curve of Figure 6 becomes greatly straightened when the power ratio is taken as the

measure of signal strength, which in turn seems to show that the detector is not losing its efficiency as fast as at first sight appears with decreasing signal strength.

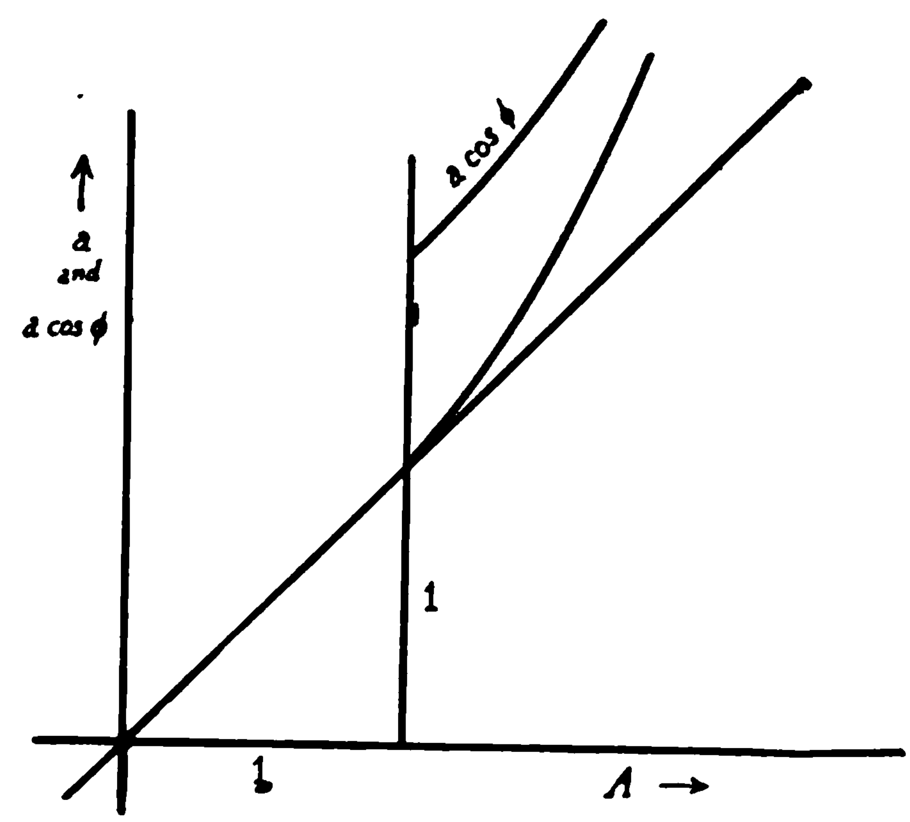

FIGURE 8

The practical importance of the measurement of signal strength arises out of the fact that for purposes of design the law giving strength of signal with distance of propagation should be accurately known. At the present time we have to use the Austin-Cohen formula, which is apparently based on observations of the quantity A'. The records could be reduced to terms of the power ratio $a \cos \phi$ by means of the formulas and diagrams above if Austin and Hogan had made it clear which of their various definitions had been used on the various occasions and had fully chronicled the relevant details of their apparatus.

The difficulties attending the use of the audibility method of measurement of signal strength, including the prime difficulty of the uncertainty of the behavior of the crystal detector, may all be overcome by calibrating the detector and circuits repeatedly, or, what is perhaps better still, by measuring the strength of the received signals by balancing their telephone sound against locally produced signals of the same pitch and of adjustable measurable intensity.

A number of circuits for doing this have been proposed for damped waves, and some that have been used by the author and found trustworthy at sea are shown in Figures 9, 10, and 11. They were used with damped waves, but can be used with chopped continuous waves without change. The method con-

sists in exciting the antenna by means of a tuned circuit giving very feeble oscillations of adjustable strength, which is coupled to the antenna to an extent decided upon beforehand as suitable. These locally produced signals are of course heard in the telephone at the same time as the signals from a distance and can be adjusted to the same audibility. The local oscillations are pro-

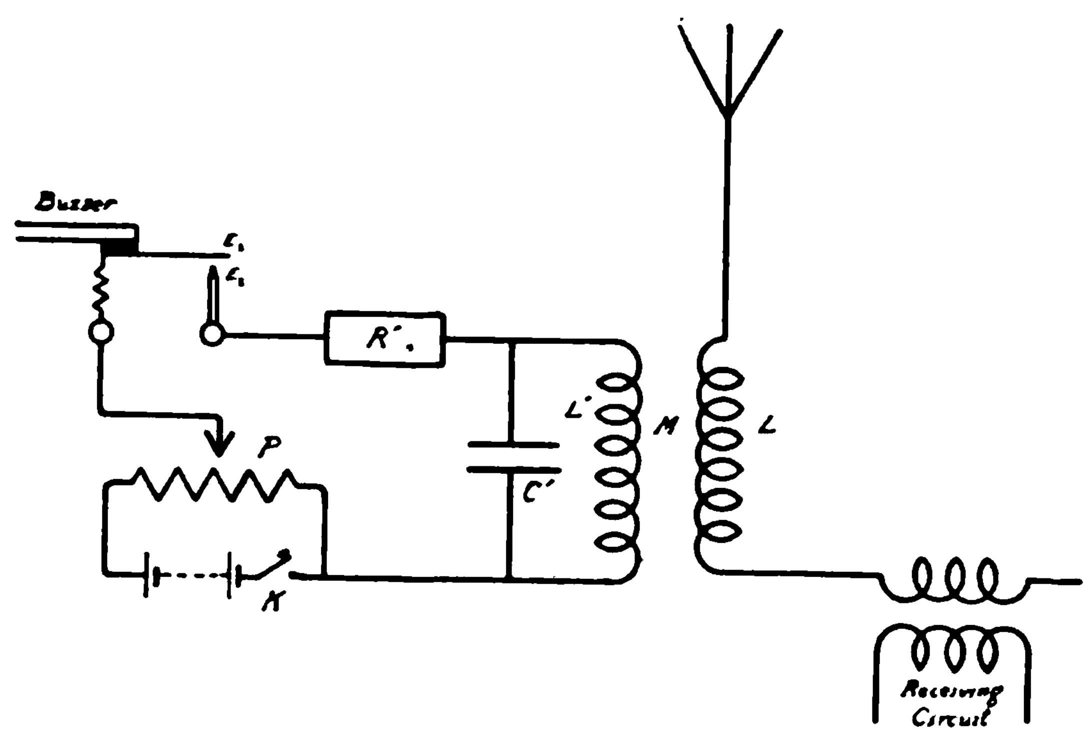

FIGURE 9

duced in the circuit $L'C'$ by means of the contact $E_1 E_2$, which is rapidly made and broken by the buzzer B and thru which, when contact is made adjustable, electromotive force is applied from the potential divider P. The beating contact E_1 E_2 must be carefully insulated from the buzzer, and the magnet of the buzzer is partly short-circuited by a resistance to prevent sparking at the driving contact breaker. The coil L' is coupled to the an-

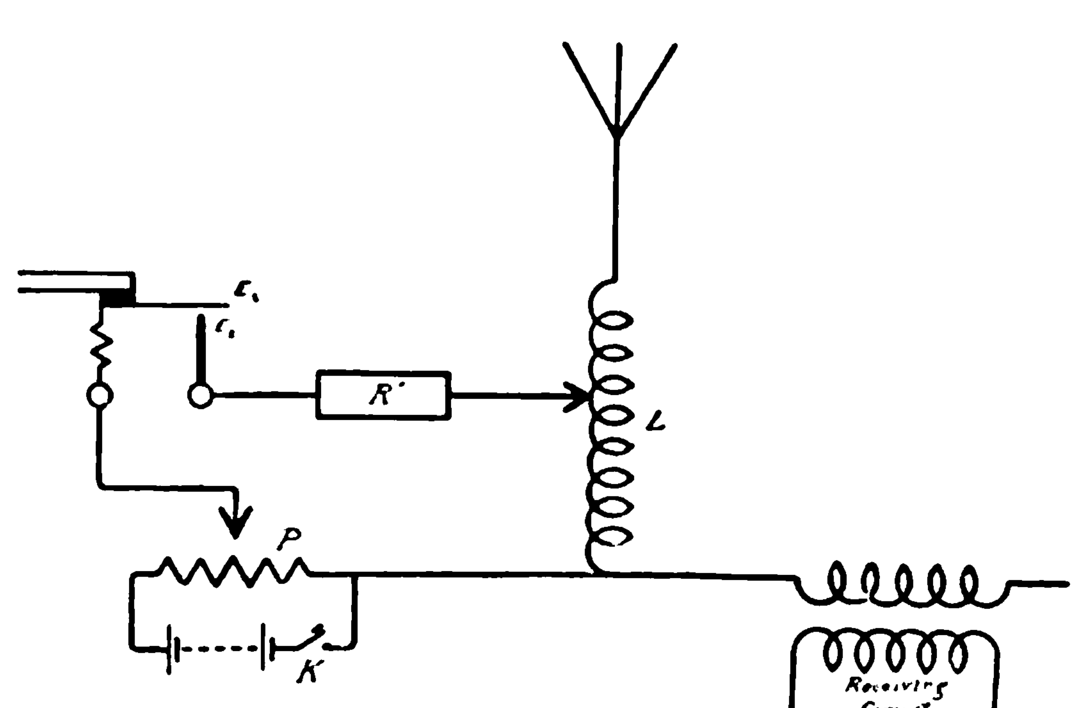

FIGURE 10

tenna loading coil L, and M is set before the experiments at a convenient value. The key K in the circuit of the potential divider enables the operator to send dots and dashes. The buzzer should be adjustable in pitch and made to give the same note as the signals being made. The circuit $L'C'$ must be tuned

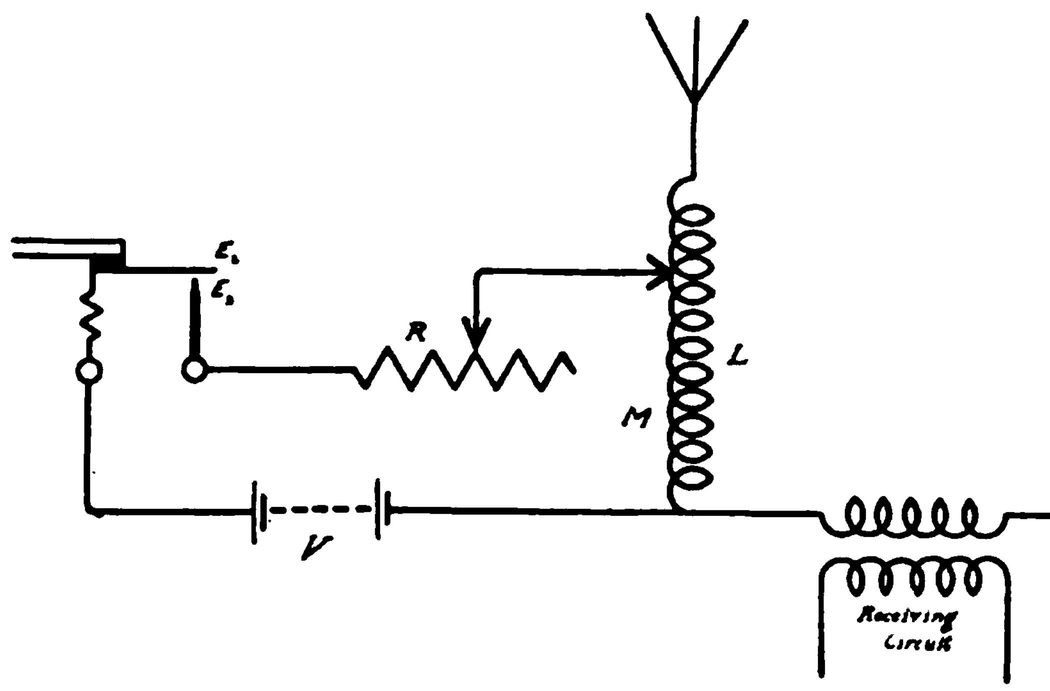

FIGURE 11

to the antenna and to the waves. Ideally the damping of this circuit should be made the same as that of the signals received. Instead of the tuned circuit $L'C'$ a plain inductance coil may be used; that is to say, the condenser C' could be removed; or, even simpler, L' may be made to coincide with a portion of L, as indicated in Figure 10. In this case also it is well to insert a large resistance R' in the circuit of the potential divider. The antenna is in this mode of operation said to be excited by "impulsing." A simpler circuit still is shown in Figure 11, where R is a high resistance consisting of an electrolyte in a glass tube with one moveable electrode. With this last apparatus the signal strength is matched by varying the high resistance. All these methods gave, during such trials as the author could carry out on commercial stations working their ordinary programmes, practically the same measures, and appeared easier to operate with confidence than the shunted telephone method.

The above three methods are sketched in order to show how very simple the apparatus may be for improving upon the unassisted audibility method. So far as the author's experience goes with, for instance, the last of the three, the possible accuracy is much greater than with the audibility method. But it should be mentioned that in truth no really satisfactory comparison between the diffraction theory of the propagation of waves round

the globe and the experimental facts can ever be obtained with trains of damped waves, if only for the reason that the theory has regard only to sustained waves. Thus it is to be hoped that future experimenters will use sustained waves, which, besides being nearer to theoretical conditions, will enable all sorts of difficulties arising from the unknown behavior of detectors and from the presence of audio harmonics to be eluded. It is possible also, by the aid of sustained waves, to escape all the doubts which arise from the use of that crude measuring instrument, the telephone receiver. The author has recently been conducting laboratory measurements in which oscillations were received, heterodyned, rectified, the low frequency results amplified, and then measured by a tuned vibration galvanometer, and has found that conditions can be kept very steady even with high amplifications. The necessary modifications of the damped oscillation circuits of Figure 9 are obvious; the buzzer must be replaced by an ionic relay so as to sustain oscillations in the circuit $L'C'$ and the intensity of the oscillations induced in the antenna for heterodyning must be varied either by varying M or varying the intensity of the oscillations in $L'C'$. The receiving circuit would of course be one adapted for beat reception.

SUMMARY: The "shunted telephone" method of measuring audibility of received signals is discussed, and it is shown that the audibility factor as usually calculated, may vary widely from the true strength ratio. This is true whether shunted resistance or shunted impedance is taken as the basis of calculation.

The author then determines the radio quantity corresponding to any determined audibility factor or strength ratio. This is of importance in connection with quantitative measurements on long distance transmission.

An alternative comparison method of measuring incoming signal strength is described, wherein the antenna may be excited from a local buzzer of adjustable pitch and having an independent contact for the antenna "impulsing circuit." This method is regarded as more accurate than the usual one.

Sustained waves should be used for transmission experiments, and these may be received, heterodyned, rectified, amplified, and measured quantitatively by a vibration galvanometer.

DISCUSSION

Louis W. Austin (by letter): Doctor Eccles' paper is of great interest to me on account of the long use of the shunted telephone method in my laboratory. His Figure 5 shows that the audibility, when the impedance of the telephones is used in the calculations, is practically proportional to the strength ratio, except for the curved portion of the hyperbola lying to the left of unit audibility, where, of course, it has no physical meaning, as there Z would be greater than infinity.

On account of the difficulty in determining the effect of the shunt with high frequency oscillations, I always prefer, in my work, to calibrate the apparatus by a comparison of the telephone shunt readings with the readings of a galvanometer, which can be connected in place of the telephones. In the case of silicon, perikon, and many other crystal detectors, the galvanometer readings are strictly proportional to the squares of the radio frequency currents. The method used in the calibration for oscillating audion reception have already been described to the Institute (PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 5, page 239, 1917).

In one place in his paper, Doctor Eccles refers to uncertainty in regard to whether resistance or impedance was used in the calculations of the Brant Rock experiments. ("Bulletin of the Bureau of Standards," volume 7, page 315, 1911.) Professor Love has assumed ("Transactions Royal Society," London, Series A, volume, 215, page 105, 1915) that the resistance of the telephone, 600 ohms, not the impedance, 2,000 ohms, was used, and has used his conclusions to support the MacDonald transmission theory. I have never been able to understand how such an uncertainty could have arisen, as the question of resistance or impedance does not enter into the plan of calibration used, which is shown on page 319 of the paper and is explained in the text. There Table 1 gives the currents measured in the antenna with the silicon detector "D" and the corresponding shunts required to reduce the telephone currents to audibility. As has been said, the deflections of a galvanometer with a silicon or perikon detector are proportional to the squares of the radio frequency currents. Table 1 was made from the smooth curve giving the relation of telephone shunt to observed antenna current. The values given in the table show that using the impedance of the telephone, 2,000 ohms (note 8, page 318 of the paper) the antenna

current is approximately proportional to the square root of the audibility calculated from the equation $A = \frac{Z+S}{S}$.

I have made many experiments with Doctor Eccles' comparison methods, examples of which are shown in his Figures 9, 10, and 11, but cannot say that I have found them better than the shunted telephone. For several months last year, experiments were carried on in matching the beat tone in telephones used in oscillating audion reception, by means of a known variable current of the same frequency in a circuit into which the telephones could be connected. This was practically the same arrangement recently described before the Institute by Doctor Van der Bijl in his paper on detecting efficiency of the thermionic detector. This is, undoubtedly, the most accurate method of measuring telephone currents in the laboratory, the results agreeing within two or three per cent. But unfortunately when atmospheric disturbances are present, as is usually the case in long wave reception, their presence in one circuit and absence in the other, renders the accuracy of measurement no greater than with the shunted telephone.

THE CABOT CONVERTER*

By

CLAUDE F. CAIRNS

During the years 1912 to 1915, the writer had the pleasure of working with Mr. Sewall Cabot of Brookline, Massachusetts on the development of a polyphase commutator, or machine for converting polyphase alternating currents of commercial voltages to non-fluctuating voltages as high as 100,000. During that time machines were built to deliver voltages within these limits, some of which are still in commercial operation.

The field for the Cabot converter is that covered by all machines now employed for the production of direct current of any desired voltage from an alternating current or direct current source of another voltage, and furthermore as direct current generating machinery is unsatisfactory for voltages over 2,000, the Cabot converter fulfills a long felt want in efficiently supplying voltages greater than this.

The Cabot converter consists of a constant potential polyphase transformer with a primary winding exciting a relatively small number of magnetic circuits, and a secondary winding of a relatively large number of phases, from which wires are led to commutating parts driven by a motor in synchronism with the alternating current supply. The transformer is the essential part of the apparatus, as it is here that the secret of the successful operation lies. The primary is usually wound as a three-phase delta-connected winding directly connected to the alternating current source, and the secondary is usually a nine-phase ring-connected winding. Altho the primary may be wound for any number of phases greater than one and the secondary any number greater than nine, as will later be seen, three-phase to nine-phase winding readily lends itself to good mechanical and electrical practice. The method of obtaining a nine-phase ring winding can best be shown diagrammatically. Figure 1 indicates the method.

The six-sided symmetrical polygon shown in Figure 1 has

* Received by the Editor, November 26, 1917.

adjacent sides the ratio of which are 1.79 to 1.49, and which are 120° apart. A circle with its center at the center of the polygon may be drawn which will cut the sides in nine points equally spaced or 40° apart. It can readily be seen that if the number of

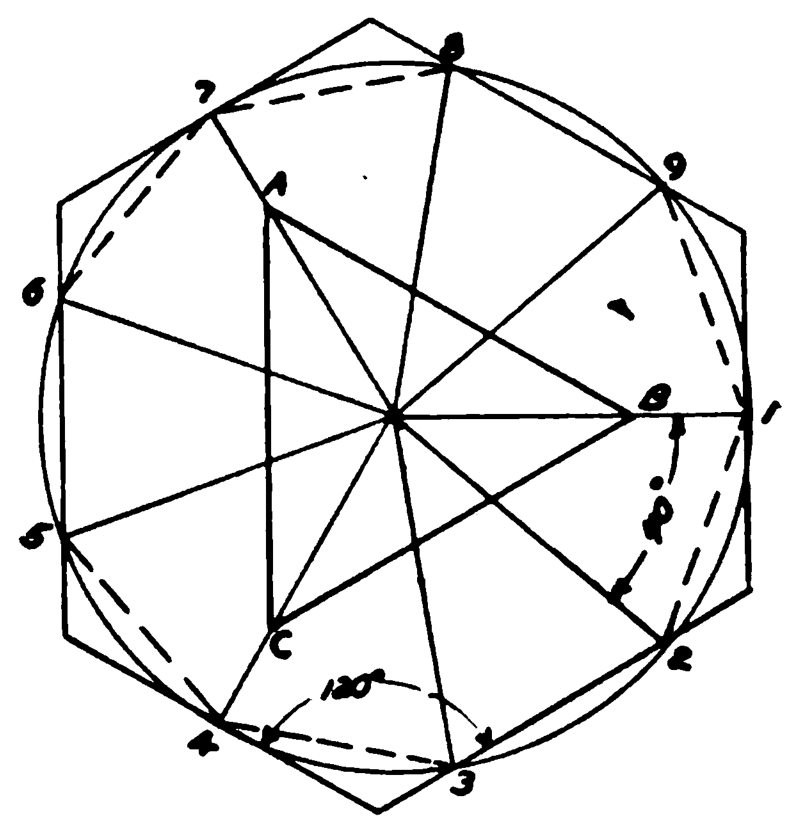

FIGURE 1—Diagram of 3-Phase to 9-Phase Transformer

turns used in the secondary winding are such that their ratio is the same as that of the symmetrical polygon, if they are connected in the proper direction, and if leads are brought out at the proper points, a ring winding will be formed the same as with a Gramme ring having nine phases. Inside of the diagram is shown the three-phase delta-connected primary winding marked A, B, C. Three of the nine phases are made up of windings which are in phase with the primary windings and the other six phases are made up of two windings connected together, one of which is in phase with one of the primary phases and the other in phase, or rather 180° out of phase, with another primary phase. If the turns of one of the nine phases totally in phase with the primary is taken as 1, then the turns of the windings which make up the other phases will be 0.395 and 0.743. The secondary of a three-phase to nine-phase transformer has therefore inherently about fifteen per cent. more copper than a straight three-phase to three-phase transformer, for in order to get a given voltage for six of the phases it is necessary to use two windings of unequal voltages, 120° out of phase.

The leads from the nine points equally spaced, which means

equal voltage between them, are connected to the commutating parts just as are the leads of an ordinary armature winding. When using a four-pole motor for driving, the commutating parts consist of nine brushes on nine slip rings, an eighteen-segment commutator, and four brushes, for low voltage machines, or nine brushes equally spaced around a four-segment commutator and two slip rings for high voltage machines. In the case of low voltage machines, the voltage between segments is that of each phase or $\frac{1}{4.05}$ of the total d. c. voltage; and in the case of high voltage machines is the vector sum of the voltages of two phases, as every other phase lead is connected to adjacent brushes, and the opposite segments are connected to the same slip ring.

In all machinery used for producing non-fluctuating direct current, the practice is to have a stationary electrical field and to rotate the iron and wire. The Cabot converter, however, has a rotating electrical field and holds the iron and wire stationary. The problems of commutation are very similar except that the Cabot converter has no armature reaction and for this reason readily lends itself to mathematical treatment and allows the use of a higher voltage between segments on the commutator. As in d. c. armature construction, the load current of the Cabot converter passes into and out of the secondary of the transformer at opposite points, and divides equally thru the two paths of the system. For successful commutation, the current flowing in any one of the phases must come to zero and reach an equal magnitude in the opposite direction during the time such phase is kept short circuited by the brushes on the commutator. As is well known, the inductance of the phase keeps the current flowing in the same direction, this necessitating an e. m. f. being built up in the opposite direction in order to bring the current to zero, and to establish it in the opposite direction. The force tending to keep the current flowing in the coil is known as the reactance voltage and is specifically equal to the inductance times the rate of change of current: $L\frac{di}{dt}$. As the current falls along practically a straight line, this can be written $L\frac{I}{t}$, where I is one-half the full load current and t is the time of short circuit. In all previous work on commutation of stationary polyphase windings, it was without doubt felt that the inductance of the phase commutated is the total inductance of the winding. On the contrary, how-

ever, it is the leakage inductance of the phase which makes itself felt at the commutator.

The transformer is therefore wound so that the leakage inductance of the phases shall be a minimum consistent with good practice. This is accomplished by having each of the phases and each of the windings used in making up a phase of the secondary linked thruout their winding length with a complete primary winding. By so winding the transformer, it is possible to employ the well-known formula for leakage inductance:

$$L = \frac{4\,\pi\,N^2\,p}{l}\left(\frac{x}{3}+\frac{y}{3}+g\right)10^{-9} \text{ henries.}$$

This applies to a two layer winding on a core type transformer, where

N = number of turns in phase or winding,
p = mean turn in cms.,
l = length of winding channel in cms.,
x = radial depth of primary in cms.,
y = radial depth of phase or winding in cms.,
g = width of the gap between primary and phase or winding in cms.

Of course, by further sandwiching the coils, the leakage inductance can be further reduced.

The leakage inductance of a transformer so wound is very much less per turn for a given output than in armature construction which necessitates an air gap; and, as mentioned before, there is no such phenomena as armature reaction with which to contend. This readily explains why successful commutation can easily be obtained with only nine phases with the Cabot converter against sixty or more with armature construction, without employing any electrical or mechanical means of producing a commutating e. m. f.—other than the resistance of the circuit—or shifting the angle of commutation. For low voltage machines, up to 220 volts, the leakage inductance of the transformer is so small that very low resistance brushes may be employed, thus effecting a great saving in the size of the commutator and the watts lost due to friction and $I^2 R$ losses.

When, however, the d. c. voltage to be delivered attains higher values, the leakage inductance of the transformer increases about proportionately, and means must be provided for shifting the angle of commutation in order to utilize the e. m. f. induced in the winding, or for providing some form of commutating e. m. f. successfully to reverse the current. If the brushes

are given a permanent forward shift so as to have sparkless commutation at full load, sparking will occur at no load and vice versa. The amount of necessary shift can, however, be readily calculated by a simple mathematical process.

Consider the circuit shown in Figure 2.

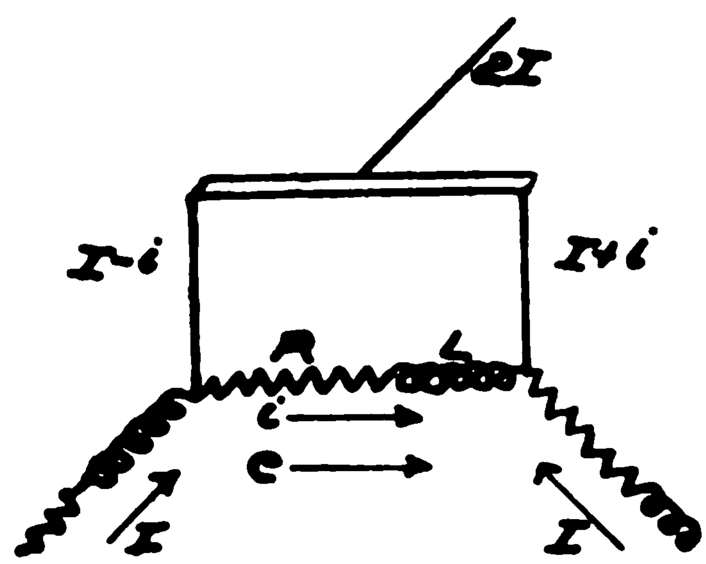

FIGURE 2

The resistance of the brush has been neglected, as in high voltage machines it is negligible compared with the winding.

L = the leakage inductance of the phase under commutation,
R = the resistance of that phase,
I = the current flowing in the ring or one half the load current,
i = the current in the phase at any instant,
e = the e. m. f. at any instant induced in the phase.

Then
$$L\frac{di}{dt}+Ri+e=0,$$
$$L\frac{di}{dt}+Ri+E\sin(\omega t+\phi)=0,$$
which, when solved, gives
$$i=\left[I+\frac{E}{\sqrt{R^2+\omega^2L^2}}\sin(\phi-\theta)\right]\varepsilon^{-\frac{Rx}{360fL}}-\frac{E}{\sqrt{R^2+\omega^2L^2}}\sin(x+\phi-\theta),$$
where
$$\theta=\tan^{-1}\frac{\omega L}{R},$$
ϕ = the angle at which the short circuit comes on,
x = electrical degrees of duration of the short circuit,
f = frequency of alternating current supply,
E = maximum e. m. f. induced in the phase,
$\omega=2\pi f$.

This equation gives the value of the current in the coil being commutated at any instant after the short circuit comes on, provided we know the other quantities. All the quantities are constants of the transformer except ϕ and x. The quantity x is the number of degrees the short circuit has been on at the instant at which the value of i is to be determined. Therefore only ϕ is left as an unknown which it is necessary to determine. In order to have commutation complete at the instant that the short circuit is removed, i must be equal to $-I$. With this value of i, putting x equal to the number of degrees of short circuit, ϕ can be determined.

Then the equation will be

$$-I=\left[I+\frac{E}{\sqrt{R^2+\omega^2L^2}}sin(\phi-\theta)\right]\varepsilon^{-\frac{Rx}{360fL}}-\frac{E}{\sqrt{R^2+\omega^2L^2}}sin(x+\phi-\theta).$$

I is one half the load current, and therefore for different values of I it will be necessary to have the short circuit come on at different times in order that the current may be completely commutated. Now ϕ is the angle at which the short circuit comes on, and the difference between the value of ϕ when $I=0$ and $I=\frac{1}{2}$ of the full load current is the necessary angular shift, to have sparkless commutation from no load to full load.

It is interesting to note that if intermediate values of the necessary shift between full load and no load be computed, they will be directly proportional to the load current within a very considerable degree of precision.

From the above it is evident that the requisite shift may be obtained by any leading quadrature e. m. f. which is proportional to the d. c. load or the brushes themselves may be mechanically moved forward or the commutator retarded a predetermined amount proportional to the load.

To produce a quadrature e. m. f., there could be inserted in the leads to the transformer three series transformers across the secondaries of which are connected condensers. As the load comes on, the voltage across the condensers increases, and hence the voltage impressed on the transformer is shifted forward in regard to time in proportion to the load, since the capacity reacts thru to the line as the square of the ratio of transformation. The same results may be obtained by using synchronous machinery. Either on the same shaft with the commutating parts, or driven by a separate motor, is a rotor having a field winding thru which

is passed the d. c. load current, or a part of it, by means of a shunt. Surrounding the rotor is a stator on which is a three-phase winding left open in order that the three-phase a. c. leads to the transformer may be connected in series. In this way an e. m. f. is induced in the main leads. The magnitude of this e. m. f. is determined by the field strength, and the time phase is determined by the position of the field relatively to the stator. In this way the angle of the impressed e. m. f. may be shifted proportionately with the load and in the right direction to produce a commutating e. m. f. In addition to this feature, compounding is obtained, for as the load increases the magnitude of the induced e. m. f. increases, which increases the impressed e. m. f. and also changes its angle.

To shift the brushes forward mechanically, or to do what amounts to the same thing, that is, to retard the commutator, the best method is shown in Figure 3. A copper disc, mounted on the shaft with the commutator, is made to rotate in a magnetic field produced by a winding in series with the d. c. load current. As the load current increases the eddy currents induced in the disc retard the rotor of the driving motor and change its running angle by an amount proportional to the d. c. load.

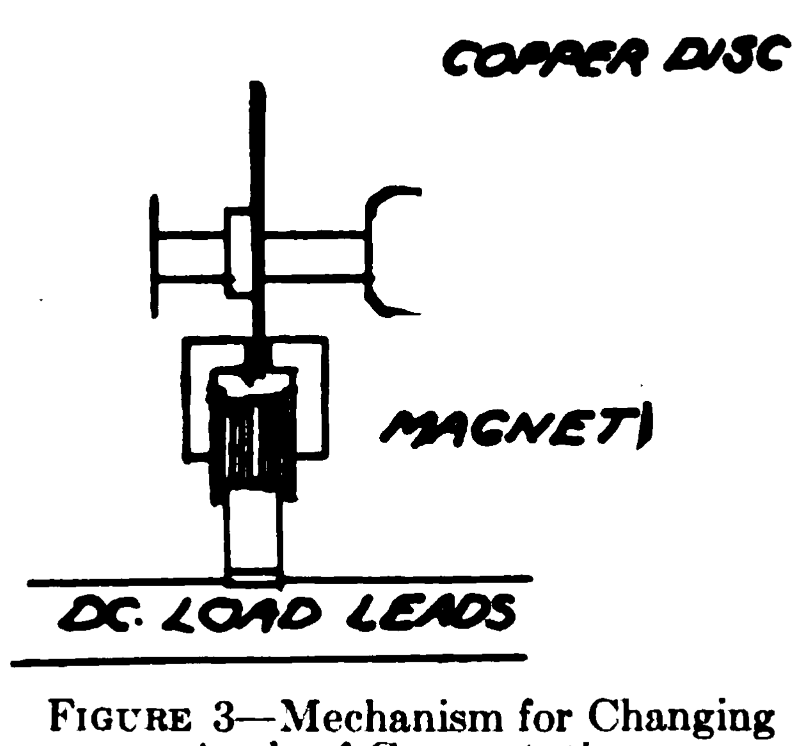

FIGURE 3—Mechanism for Changing Angle of Commutation

The same results can be obtained by introducing resistance into the circuit in which the commutating current is at any instant flowing. By this method the current flowing in the short-circuited coil is brought to zero by dissipating itself in heat and is established in the opposite direction by the e. m. f. induced in the same coil by the primary. When a four pole machine is used

for driving, the commutating parts consist of nine brushes, a four-segment commutator, and two slip rings. This construction readily lends itself to a form whereby the same resistance may be employed for all phases, thereby eliminating the necessity of putting resistance in each phase or brush. The two opposite segments on the commutator are connected to the same slip ring and since the phase being commutated at any instant is short-circuited by these two segments and slip ring, resistance may be inserted in the wires connecting these three in series. The amount of resistance necessary can readily be calculated by the same method as that employed in calculating the necessary angular shift for complete commutation. To do this, it is necessary to find what value of R in the previous formula will give practically the same value for ϕ when $I=0$ and when $I=$one-half the full load current. In determining the value of R to insert in the wires, the cut-and-try method very often serves very well, as the angle of commutation is not very critical and a slight excess is detrimental only to efficiency.

When the d. c. voltage reaches values as high as 80,000 to 100,000, the leakage inductance of the transformer attains such values that the necessary shift in the angle of commutation becomes very appreciable. Altho a complete investigation has never been made it was found that with such voltages the use of what has been termed serial subdivisions gave satisfactory results for low power. The serial subdivisions consist of brushes and segments so timed by their position on the shaft that when the circuit of the coil being commutated is broken the break occurs at several points simultaneously. This procedure brutally breaks the current which at that time may still be flowing in the circuit, and has no practical value for larger powers than a few kilowatts.

Altho machines of large power output have never been built, the design of such machines show that the problems in commutation are the same as for small power, and that the reactance voltage is independent of power for a given voltage, since the increase in current is offset by the decrease in the leakage inductance of the windings. In low-voltage machines of high power the inductance of the leads from the transformer to the commutator, will, unless properly grouped, be in excess of the leakage inductance of the winding.

The production of non-fluctuating direct current from alternating current by the rotation of iron and wire is accompanied by losses in electrical energy greatly in excess of the copper losses in

the armature, and the active material used for magnetization purposes is nearly three times the amount used in the armature for conversion purposes, because of the necessary air gap. The Cabot converter, however, accomplishes the desired result with a relatively large efficiency as the losses are only those of the transformer, the brush I^2R, the input into the motor used to overcome the brush friction, and the losses (in high voltage machines) due to the method used for shifting the angle of commutation or for providing a commutating e. m. f.

The losses in the transformer are of course small. In low-voltage machines, the reactance voltage is so low that the lowest resistance carbon brush may be used, resulting in small brush I_2R and small friction watts owing to the higher current density permissible with a low resistance brush. In high voltage machines, the brush resistance becomes negligible, and the power used for aids to commutation is relatively small especially for high power output.

The space occupied by a Cabot converter is much less than that required for a rotary converter or motor generator capable of handling the same amount of power owing to the small amount of active material necessary for excitation purposes in the transformer and the small commutating parts necessary to handle the load.

The Cabot converter has already been used for various purposes where non-fluctuating direct current is needed both for high- and low-voltage work.

For low-voltage work, a machine was employed for charging storage batteries, converting 220-volt three-phase 60-cycle alternating current to 90- to 150-volt direct current, with a current value of 100 amperes. A machine has been employed for exciting a direct current arc lamp and one has been tested for its ability to run direct current motors with marked success.

For high-voltage work several machines have been in use for the excitation of X-ray tubes giving voltages up to 100,000 at 5 to 7 k. w. Some of these machines are still running today after five years of service. Machines have also been built for radio telegraph and telephone work at voltages from 900 to 2,000, to deliver in the neighborhood of 2 k. w. These machines have been adapted to spark gaps with tone circuits, but could easily be used wherever high voltages are employed in radio work.

The advantages of the Cabot converter over the other forms of apparatus for converting alternating currents to non-fluctuating direct current and machines for producing high potential

non-fluctuating direct current are very marked. For storage battery charging and general low-voltage work, the advantages are the ease with which the voltage may be varied by changing the primary turns, the greater efficiency, the lower first cost, and the smaller floor area.

In the high-voltage field, the advantages are even more marked than in the low voltage field. In X-ray work, the ability to read the true voltage across the tube together with current and the time of exposure gives the operator a direct measurement of the watts per second, and hence the dosage given to a patient. Furthermore in radiographic work results can be duplicated as the penetration or quality of X-rays is proportional to the voltage and the quantity is proportional to the current and time.

In radio work, machines with rotating iron and wire become unreliable above 2,000 volts because of commutation and the difficulties of insulation. With the Cabot converter the insulation problems are no greater than with any high potential transformer, and sparkless commutation can be obtained for even higher voltages without the use of mechanical or electrical means of shifting the angle of commutation but simply by the addition of resistance to the commutating circuit. For use on ships the Cabot converter would readily adapt itself because of the fact that any break-down could easily be attended to by the operator assuming him supplied with a small box of spare parts.

Below are given the specifications for the transformer of a 2,000-volt, 1 k. w., radio transmitter.

A. C. supply voltage: 136 volts, 3-phase, 120-cycles,
D. C. volts: 2,000,
D. C. amperes: 0.5,
Iron watts: 50,
Copper watts: 60,
Iron weight: 16 pounds (7.3 kg.),
Copper weight: 4.2 pounds (1.9 kg.),
Percentage efficiency: 90,
Temperature rise: 50°,
Reactance voltage: 15.7 volts for 12° (electrical), short circuit,
Necessary shift in the angle of commutation from no load to full load: 0.5 electrical degrees.

The transformer is wound as shown in Figure 4.

Under each secondary winding is wound a complete primary winding, all of which, on the same leg, are connected in

parallel. By so winding the transformer, each secondary winding is completely linked with a primary winding and therefore the leakage inductance formula may be applied to each section.

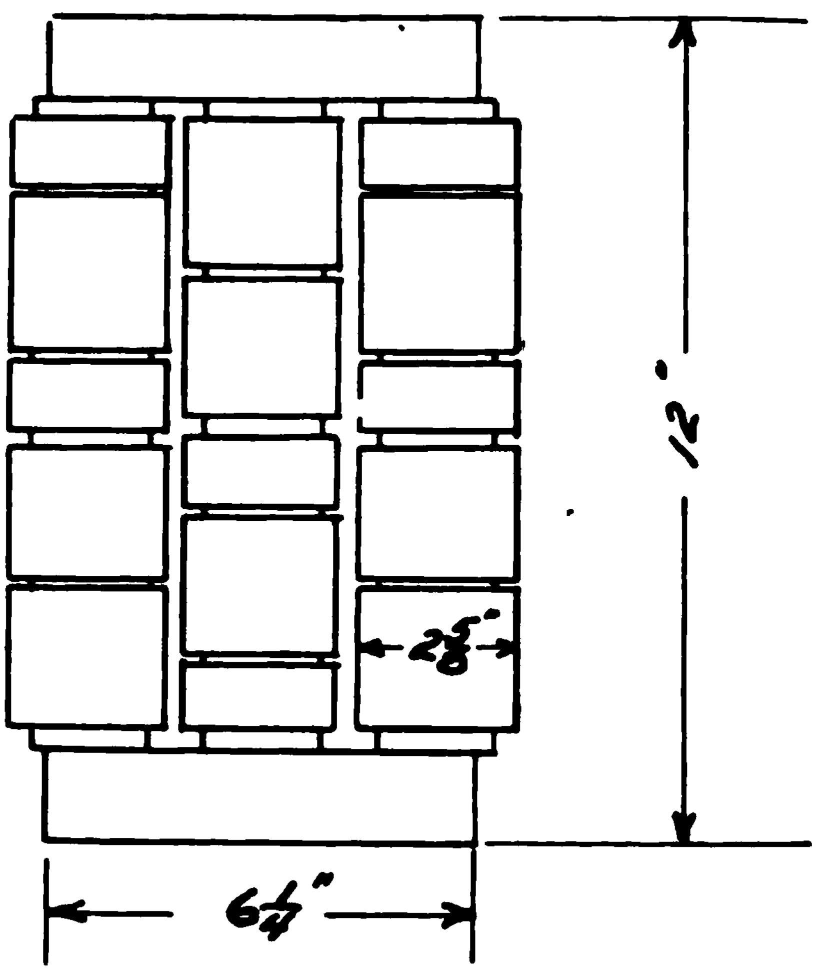

FIGURE 4—1 K. W. Transformer for 2,000 Volt D. C. Cabot Converter

Altho the angular shift is only 0.5 degrees, by the addition of a few ohms of resistance in the commutating circuit the necessary shift can be reduced to zero which means that sparkless commutation will occur from no load to full load. This transformer was actually built and used.

There has been built and put into operation a 5 k. w. machine to convert 220-volt, 3-phase, a. c. to 110-volt d. c. from which a 10 k. w. load was taken for 24 hours and a 15 k. w. load for several minutes with sparkless commutation! The reactance voltage was only 0.356 volts against a reactance voltage of 2 to 3 volts for a standard rotary converter of 5 k. w. output. The over-all efficiency of this apparatus was 90 per cent. at full load with a very flat efficiency curve.

SUMMARY: The Cabot converter is a combination of a few-phase primary and many-phase secondary transformer and a secondary circuit commutator driven by a synchronous motor. Direct current of high voltage can be readily produced, e. g. for radio transmitters. The converter has a lower reactance voltage than the standard rotary converter and has other constructional and electrical advantages.

ON THE POULSEN ARC AND ITS THEORY*

(SUPPLEMENTARY NOTE)†

BY

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A. MAGNITUDE OF PEAK ARC VOLTAGE REQUIRED FOR MAINTAINED OSCILLATIONS OF THE FIRST KIND

An approximate solution of this question is quite simply obtained, and may, therefore, be of some interest. Figure 1 shows the arc current, i_1, and arc voltage, e_1, in the case of sustained oscillations of the first kind. The arc voltage is supposed to be

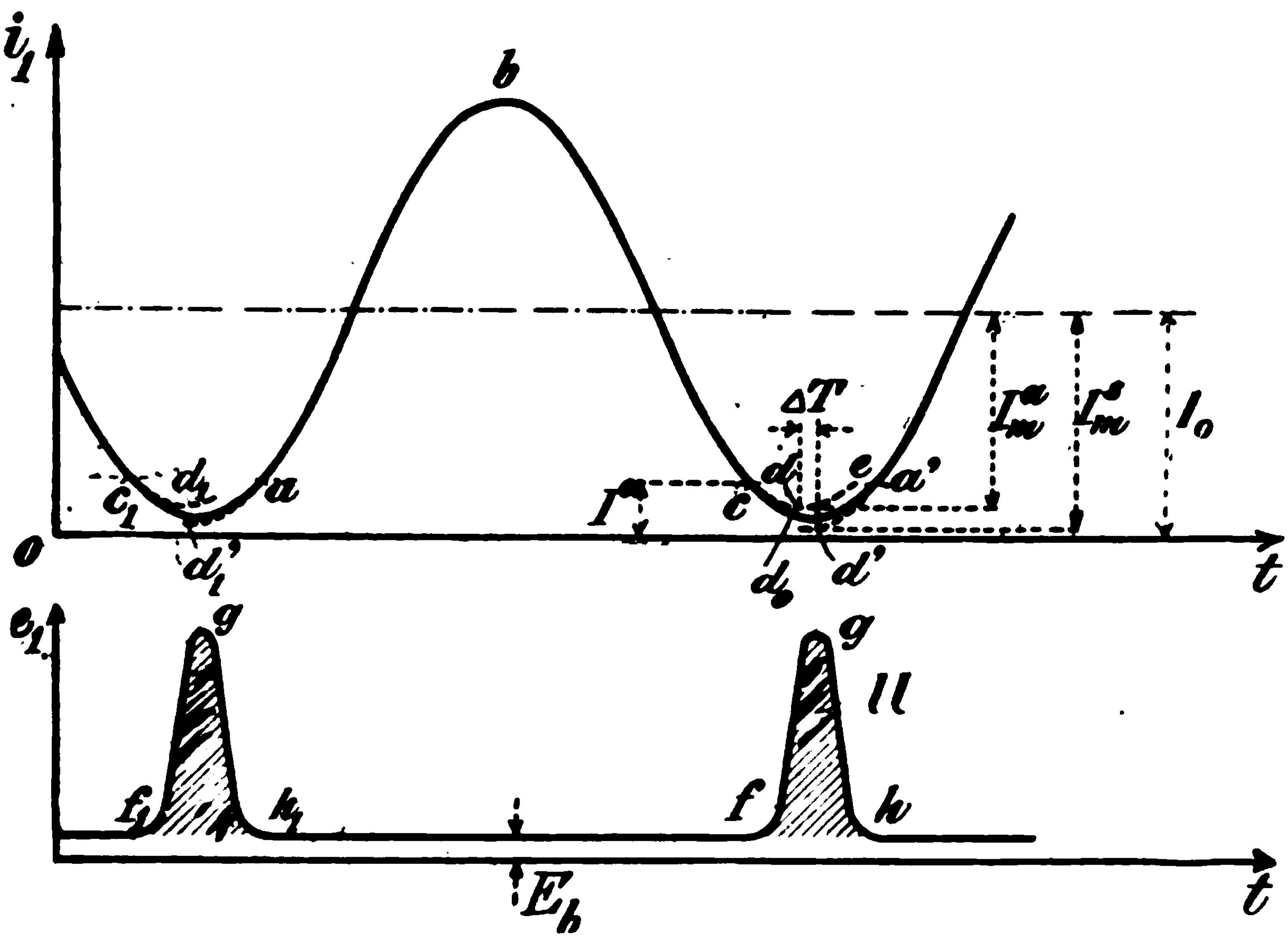

FIGURE 1—Diagram of arc current (i_1) and arc voltage (e_1) for oscillations of the first type (that is, oscillations without interruption of current thru arc)

* PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 5, number 4, page 255. The nomenclature given on pages 315 and 316 of that paper applies also to the present paper.

† Received by the Editor, December 18, 1918.

constant and equal to E_b for currents greater than I^o. The part $a\,b\,c$ of the current curve can, therefore, be written in the form

$$I_o - i_1 = i_{10} = I_m^s\,\varepsilon^{-\kappa t}\cos(2\pi n t - \phi), \tag{1}$$

where I_m^s is the initial maximum amplitude of the current; while

$$\kappa = \frac{R}{2L}, \text{ and } 2\pi n\sqrt{LC} = 1.$$

At the point c of the current curve, the voltage commences to increase, and $f\,g\,h$ represents the peak; the corresponding part of the current curve is $c\,d_o\,a'$. We shall now determine the relation between the curves $f\,g\,h$ and $c\,d_o\,a'$.

For the arc voltage e_1 we have

$$-e_1 = L\frac{d\,i}{d\,t} + R\,i + \frac{1}{C}\int i\,d\,t, \tag{2}$$

i_1 being the arc current and L, R, and C the constants of the r.f. circuit.

If we put $i = i_{10} - i_1'$ where i_{10} is a current as determined by (1), then i_1' is the difference between this current and the actual arc current, that is, between the curves $c\,d\,e$ and $c\,d_o\,a'$ in Figure 1.

Substituting the above expression for i_1 in (2), we get the following equation:

$$e_1' = e_1 - E_b = L\frac{d\,i_1'}{d\,t} + R\,i_1' + \frac{1}{C}\int i_1' d\,t. \tag{3}$$

However, i_1' is so small that the last two terms may be neglected, and (3) may therefore be written

$$e_1' = e_1 - E_b = L\frac{d\,i_1'}{d\,t} \tag{4}$$

or

$$i_1' = \frac{1}{L}\int_f^t e_1'\,d\,t = \frac{1}{L}\int_f^t (e_1 - E_b)\,d\,t. \tag{5}$$

At the point h, the arc voltage again reaches its normal value E_b, and at the corresponding point a' the current curve again begins to be represented by an equation of the form shown in (1). The maximum amplitude in the point d is reduced to I_m^s, and if the oscillations are to continue with the same intensity, the maximum amplitude at the point d' must be equal to I_m^a. The peak $f\,g\,h$ must, therefore, be such as to be able just to cause this increase in maximum amplitude from I_m^a to I_m^s, if the oscillations are to continue. If the peak is lower, the amplitude will

decrease continually and the arc becomes "inactive" or non-oscillatory. If the peak is higher, the amplitude will increase, and eventually becomes greater than I_o, the oscillations then being of the second kind.

The area U of the peak such, that it is just possible to sustain the oscillations is, according to (5), determined by

$$U=\int_f^h (e_1-E_b)\,dt=Li_{1h}', \tag{6}$$

where i_{1h}' is the greatest value of i_1' and corresponds to the point a' of the current curve. For oscillations of constant amplitude, this value i_{1h}' must very nearly be equal to $I_m^s-I_m^a=\delta I_m^s$, where the logarithmic decrement $\delta=\pi R\sqrt{\frac{C}{L}}$. Equation (6) may, therefore, be written

$$U=L\delta I_m^s=\pi R\sqrt{LC}\cdot I_m^s=\frac{\lambda R}{6\cdot 10^{10}}\cdot I_m^s. \tag{7}$$

This formula agrees with formula (59) of the former paper.

Formula (7) may easily be obtained in another way. During one complete period, the feeding current delivers an amount of energy A to the r.f. circuit, and to the arc, such that

$$A=\int_t^{t+\tau} I_o e_1\,dt=I_oE_b\tau+\int_f^h I_o(e_1-E_b)\,dt=I_oE_b\tau+I_oU \tag{8}$$

The first term, $I_oE_b\tau$, represents the energy spent in the arc, while the second term, I_oU, is the energy delivered to the r.f. circuit.

Accordingly we have

$$I_oU=I^2R\tau=\tfrac{1}{2}I_m^{s2}R\,2\pi\sqrt{LC}=\pi R\sqrt{LC}\,I_m^{s2} \tag{9}$$

But I_m^s is very nearly equal to I_o, and (9) accordingly reduces to

$$U=\pi R\sqrt{LC}\cdot I_m^s$$

which was to be shown.

B. Magnetic Field of the Poulsen Arc

The experimental evidence given in the former paper showing that the arc in too weak a field behaves as shown by the sketch in Figure 19 of that paper, may, perhaps, be deemed somewhat meagre. In fact, the evidence consisted only of part b of Figure 17b. Some crater oscillograms and side views which will probably be more convincing are therefore given in Figure 2 of this paper. Parts b, b' correspond to a normal field, c, c' and d, d' to fields which are too weak. A comparison with Figure 19 of the first paper shows a complete agreement.

In parts a, a' of Figure 2 the magnetic field is too strong. The crater curve in this case, however (contrary to what is the case in parts c and d of Figure 17b of the former paper), consists only of a single curve for each period. It thus appears that the crater curve in strong fields does not always split up into a number of separate parts. This being so, there is some probability

FIGURE 2—Side Views of Arc and Corresponding Crater Oscillograms
Anode to the left. $\lambda=9{,}000$ m.; $R=0$; $I_o=20$ amperes
Parts a, a'—magnetic field too strong ($V_o=120$ volts)
Parts b, b'—normal magnetic field ($V_o=60$ volts)
Parts c, c'—magnetic field too weak ($V_o=100$ volts)
Parts d, d'—magnetic field too weak ($V_o=118$ volts)

that the preliminary hypothesis relative to the behavior of the arc in excessively strong fields put forth in the former paper, and referred to in connection with Figure 18 II, is not altogether correct. The different parts of the crater curve corresponding to the time of one period may possibly be formed simultaneously, the arc consisting of separate parts which are forced outward in the same manner by the magnetic field. From the data at hand I have not yet been able to decide which of the two views is the correct one. To quote the former paper: "A full elucidation of these phenomena will, therefore, necessitate further investigation" (former citation, page 297).*

* In the bibliography given in my former paper I had unfortunately overlooked an important paper by G. Grandquist (Nova Acta Reg. Soc. Sc. Ups., Series IV, volume 1, 1907) dealing with the theory of the Duddell arc.

SUMMARY: Continuing the discussion of sustained oscillations of the first or second type produced by Poulsen arcs, as given in his earlier paper in the PROCEEDINGS, the author derives an approximate value for the peak voltage required for the maintenance of such oscillations.

He then presents further experimental data on the nature of the arc in normal or too weak or strong magnetic fields. He also considers the question of the simultaneous versus sequential production of separate arcs during a period, when the arc takes place in a very strong field.

ELECTRICAL OSCILLATIONS IN ANTENNAS AND INDUCTANCE COILS*

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CONTENTS

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* Received by the Editor, April 18 and September 10, 1918

I. Introduction

In the following paper are outlined some results of the application of the theory of circuits having uniformly distributed electrical characteristics to the electrical oscillations in antennas and inductance coils. Experimental methods are also given for determining the constants of antennas and experimental results showing the effect of imperfect dielectrics upon antenna resistance.

The theory of circuits having uniformly distributed characteristics such as cables, telephone lines, and transmission lines has been applied to antennas by a number of authors. The results of the theory do not seem to have been clearly brought out, and in fact erroneous results have at times been derived and given prominence in the literature. As an illustration, in one article the conclusion has been drawn that the familiar method of determining the capacity and inductance of antennas by the insertion of two known loading coils leads to results which are in very great error. In the following treatment it is shown that this is not true and that the method is very valuable.

Another point concerning which there seems to be considerable uncertainty is that of the effective values of the capacity, inductance and resistance of antennas. In this paper expressions are obtained for these quantities giving the values which would be suitable for an artificial antenna to represent the actual antenna at a given frequency.

The theory is applied also to the case of inductance coils with distributed capacity in which case an explanation of a well-known experimental result is obtained.

Experimental methods are given for determining the constants of antennas, the first of which is the familiar method previously mentioned. It is shown that this method in reality gives values of capacity and inductance of the antenna close to the low frequency or static values and may be corrected so as to give these values very accurately. The second method concerns the determination of the effective values of the capacity, inductance, and resistance of the antenna.

In the portion which deals with the resistance of antennas, a series of experimental results are given which explain the linear rise in resistance of antennas as the wave length is increased. It is shown that this characteristic feature of antenna resistance curves is caused by the presence of imperfect dielectric such as trees, buildings, and so on, in the field of the antenna, which causes it to behave as an absorbing condenser.

II. Circuit with Uniformly Distributed Inductance and Capacity

The theory, generally applicable to all circuits with uniformly distributed inductance and capacity, will be developed for the case of two parallel wires. The wires (Figure 1) are of length l and of low resistance. The inductance per unit length

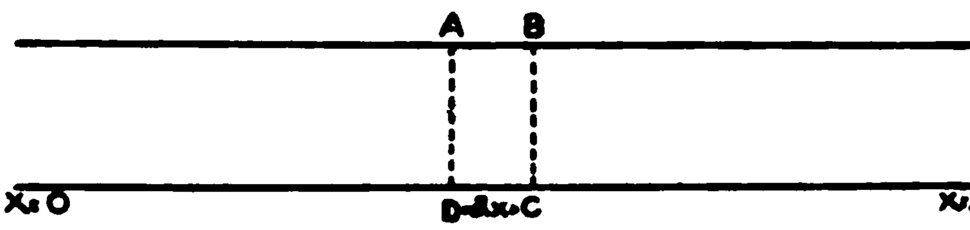

Figure 1

L_1 is defined by the flux of magnetic force between the wires per unit of length that there would be if a steady current of one ampere were flowing in opposite directions in the two wires. The capacity per unit length C_1 is defined by the charge that there would be on a unit length of one of the wires if a constant emf. of one volt were impressed between the wires. Further the quantity $L_o = l L_1$ would be the total inductance of the circuit if the current flow were the same at all parts. This would be the case if a constant or slowly alternating voltage were applied at $x=0$ and the far end $(x=l)$ short-circuited. The quantity $C_o = l C_1$ would represent the total capacity between the wires if a constant or slowly alternating voltage were applied at $x=0$ and the far end were open.

Let us assume, without defining the condition of the circuit at $x=l$, that a sinusoidal emf. of periodicity $\omega = 2\pi f$ is impressed at $x=0$ giving rise to a current of instantaneous value i at A and a voltage between A and D equal to v. At B the current will be $i + \frac{\partial i}{\partial x} dx$ and the voltage from B to C will be $v + \frac{\partial v}{\partial x} dx$.

The voltage around the rectangle $ABCD$ will be equal to the rate of decrease of the induction thru the rectangle; hence

$$\left(v + \frac{\partial v}{\partial x}\right) dx - v = -\frac{\partial}{\partial t}(L_1 i \, dx)$$

$$\frac{\partial v}{\partial x} = -L_1 \frac{\partial i}{\partial t} \qquad (1)$$

Further the rate of increase of the charge q on the elementary length of wire AB will be equal to the excess in the current flowing in at A over that flowing out at B.

Hence

$$\frac{\partial q}{\partial t}=\frac{\partial}{\partial t}(C_1 v\,d\,x)=i-\left(i+\frac{\partial i}{\partial x}\,d\,x\right)$$

$$-\frac{\partial i}{\partial x}=C_1\frac{\partial v}{\partial t} \tag{2}$$

These equations (1) and (2), determine the propagation of the current and voltage waves along the wires. In the case of sinusoidal waves, the expressions

$$v=\cos\omega t\,(A\cos\omega\sqrt{C_1 L_1}\,x+B\sin\omega\sqrt{C_1 L_1}\,x) \tag{3}$$

$$i=\sin\omega t\sqrt{\frac{C_1}{L_1}}\,(A\sin\omega\sqrt{C_1 L_1}\,x-B\cos\omega\sqrt{C_1 L_1}\,x) \tag{4}$$

are solutions of the above equations as may be verified by substitution. The quantities A and B are constants depending upon the terminal conditions. The velocity of propagation of the waves at high frequencies, is

$$V=\frac{1}{\sqrt{L_1 C_1}}.$$

III. The Antenna

1. Reactance of the Aerial-Ground Portion

The aerial-ground portion of the antenna (CD in Figure 2) will be treated as a line with uniformly distributed inductance, capacity and resistance. As is common in the treatment of

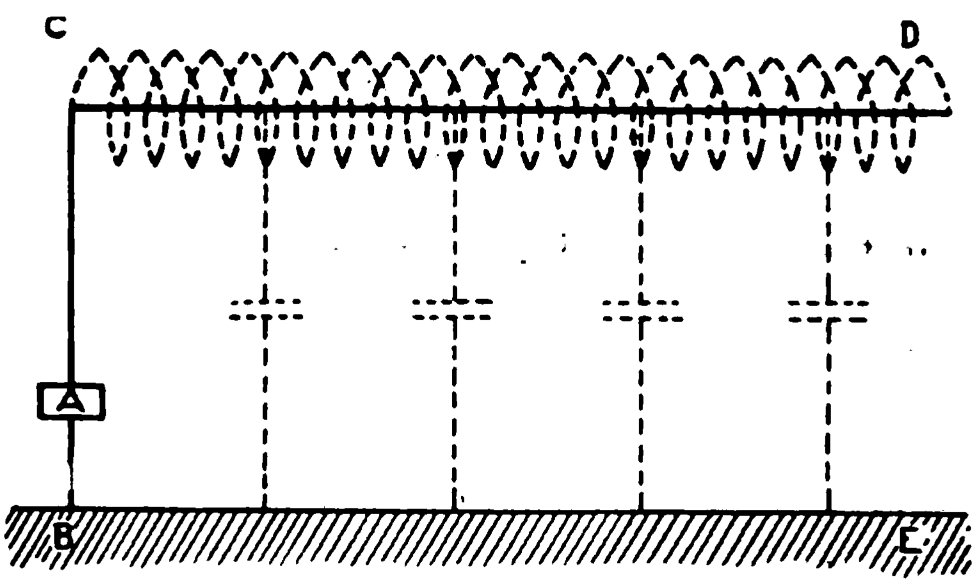

Figure 2—Antenna Represented as a Line with Uniform Distribution of Inductance and Capacity

radio circuits, the resistance will be considered to be so low as not to affect the frequency of the oscillations or the distribution of current and voltage. The lead-in BC (Figure 2) will be

considered to be free from inductance or capacity excepting as inductance coils or condensers are inserted (at A) to modify the oscillations.

Applying equations (3) and (4) to the aerial of the antenna and assuming that $x=0$ is the lead-in end while $x=l$ is the far end which is open, we may introduce the condition that the current is zero for $x=l$. From (4)

$$\frac{A}{B}=\cot \omega \sqrt{C_1 L_1}\, l \tag{5}$$

Now the reactance of the aerial, which includes all of the antenna but the lead-in, is given by the current and voltage at $x=0$. These are, from (3), (4), and (5),

$$v_o=A \cos \omega t=B \cot \omega \sqrt{C_1 L_1}\, l \cos \omega t$$

$$i_o=-\sqrt{\frac{C_1}{L_1}} B \sin \omega t$$

The current leads the voltage when the cotangent is positive, and lags when the cotangent is negative. The reactance of the aerial, given by the ratio of the maximum values of v_o to i_o is

$$X=-\sqrt{\frac{L_1}{C_1}} \cot \omega \sqrt{C_1 L_1}\, l$$

or in terms of $C_o=l C_1$ and $L_o=l L_1$

$$X=-\sqrt{\frac{L_o}{C_o}} \cot \omega \sqrt{C_o L_o} \tag{6}$$

or since $V=\frac{1}{\sqrt{L_1 C_1}}$

$$X=-L_1 V \cot \omega \sqrt{C_1 L_1}\, l$$

as given by J. S. Stone.[1]

At low frequencies the reactance is negative and hence the aerial behaves as a capacity. At the frequency $f=\frac{1}{4\sqrt{C_o L_o}}$ the reactance becomes zero and beyond this frequency is positive or inductive up to the frequency $f=\frac{1}{2\sqrt{C_o L_o}}$, at which the reactance becomes infinite. This variation of the aerial reactance with the frequency is shown by the cotangent curves in Figure 3.

[1] Stone, J. S.; "Trans. Int. Elec. Congress," St. Louis, 3, p. 555; 1904.

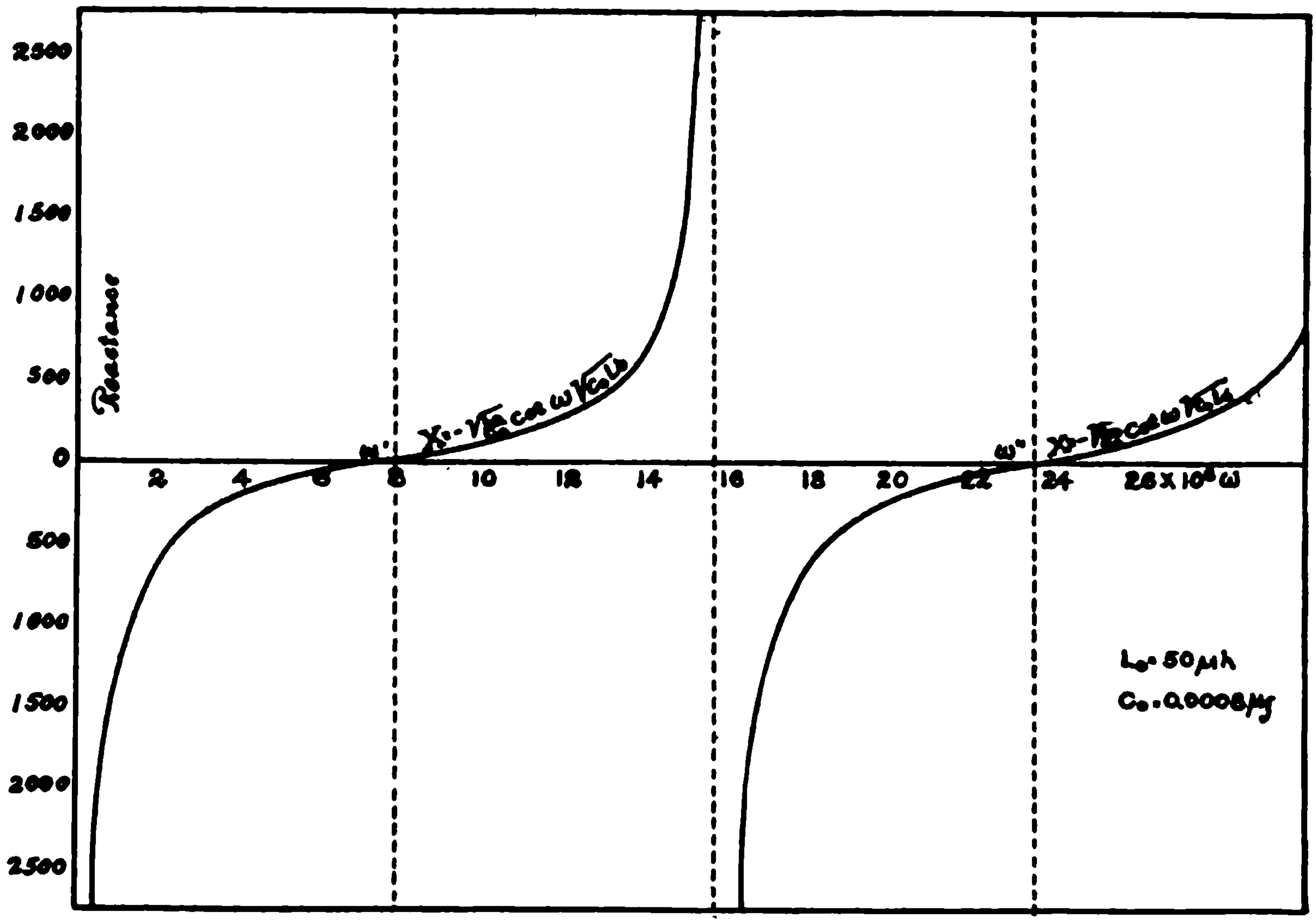

FIGURE 3—Variation of the Reactance of the Aerial of an Antenna with the Frequency

2. NATURAL FREQUENCIES OF OSCILLATION

Those frequencies at which the reactance of the aerial, as given by equation (6), becomes equal to zero are the natural frequencies of oscillation of the antenna (or frequencies of resonance) when the lead-in is of zero reactance. They are given in Figure 3 by the points of intersection of the cotangent curves with the axis of ordinates and by the equation

$$f=\frac{m}{4\sqrt{C_oL_o}};\quad m=1,\ 3,\ 5,\ \text{etc.}$$

The corresponding wave lengths are given by

$$\lambda=\frac{V}{f}=\frac{l}{f\sqrt{C_oL_o}}=\frac{4\,l}{m}$$

i. e., 4/1, 4/3, 4/5, 4/7, etc., times the length of the aerial. If, however, the lead-in has a reactance X_x, the natural frequencies of oscillation are determined by the condition that the total reactance of lead-in plus aerial shall be zero, that is:

$$X_x+X=0$$

provided the reactances are in series with the driving emf.

(a) LOADING COIL IN LEAD-IN. The most important practical case is that in which an inductance coil is inserted in the

lead-in. If the coil has an inductance L its reactance $X_L = \omega L$. This is a positive reactance increasing linearly with the frequency and represented in Figure 4 by a solid line. Those frequencies at which the reactance of the coil is equal numerically but opposite in sign to the reactance of the aerial, are the natural frequencies of oscillation of the loaded antenna since the total reactance $X_L + X = 0$. Graphically these frequencies are deter-

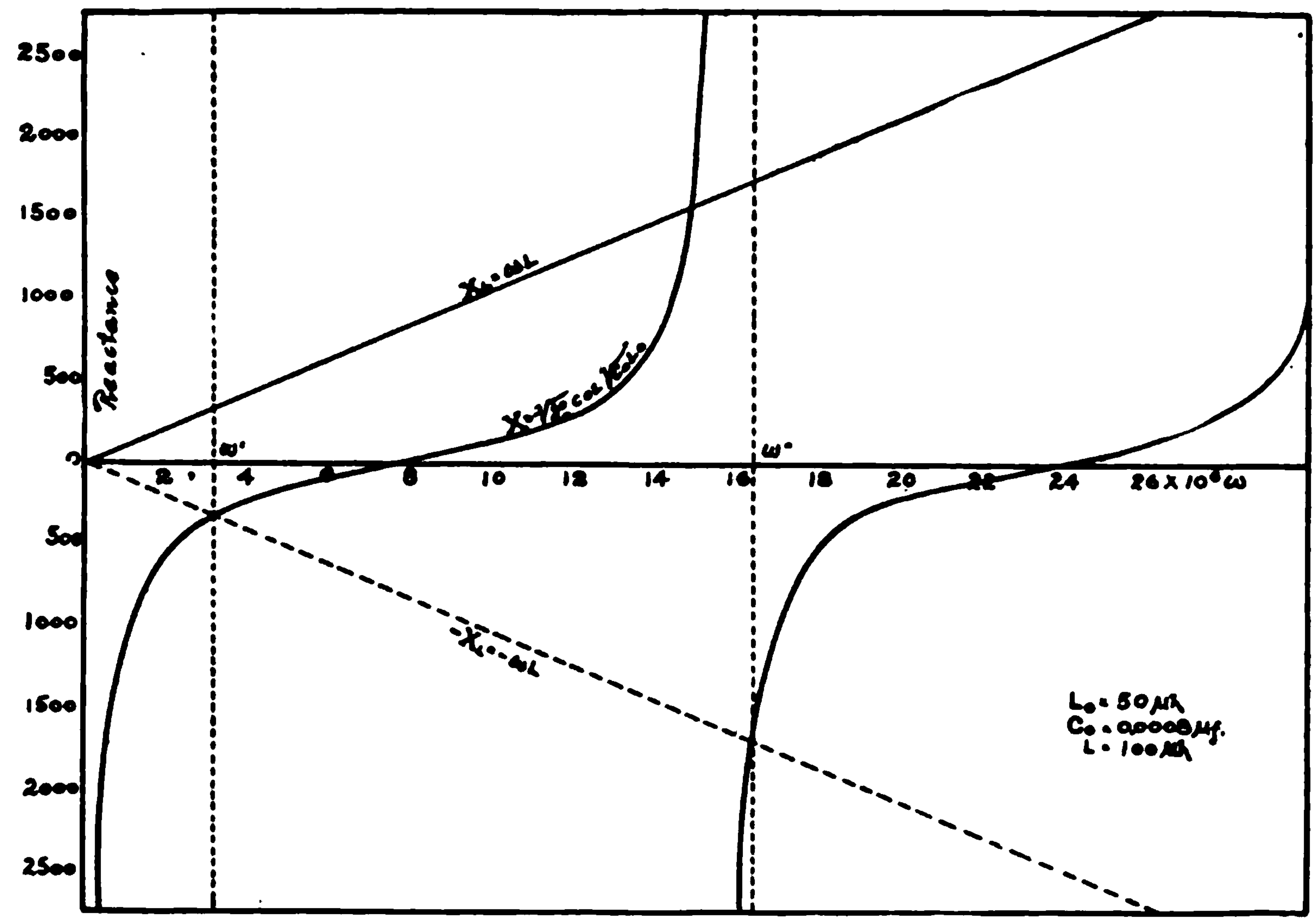

FIGURE 4—Curves of Aerial and Loading Coil Reactances

mined by the intersection of the straight line $-X_L = -\omega L$ (shown by a dashed line in Figure 4) with the cotangent curves representing X. It is evident that the frequency is lowered by the insertion of the loading coil and that the higher natural frequencies of oscillation are no longer integral multiples of the lowest frequency.

The condition $X_L + X = 0$ which determines the natural frequencies of oscillation leads to the equation

$$\omega L - \sqrt{\frac{L_0}{C_0}} \cot \omega \sqrt{C_0 L_0} = 0$$

or

$$\frac{\cot \omega \sqrt{C_0 L_0}}{\omega \sqrt{C_0 L_0}} = \frac{L}{L_0} \qquad (8)$$

This equation has been given by Guyau[2] and L. Cohen.[3] It determines the periodicity ω and hence the frequency and wave length of the possible natural modes of oscillation when the distributed capacity and inductance of the aerial and the inductance of the loading coil are known. This equation cannot, however, be solved directly; it may be solved graphically as shown in Figure 4 or a table may be prepared indirectly which gives the values of $\omega\sqrt{C_0 L_0}$ for different values of $\frac{L}{L_0}$, from which then ω, f or λ may be determined. The second column of Table I gives these values for the lowest natural frequency of oscillation, which is of major importance naturally.

(b) Condenser in Lead-In. At times, in practice, a condenser is inserted in the lead-in. If the capacity of the condenser is C, its reactance is $X_c = -\frac{1}{\omega C}$. This reactance is shown in Figure 5 by the hyperbola drawn in solid line. The intersection

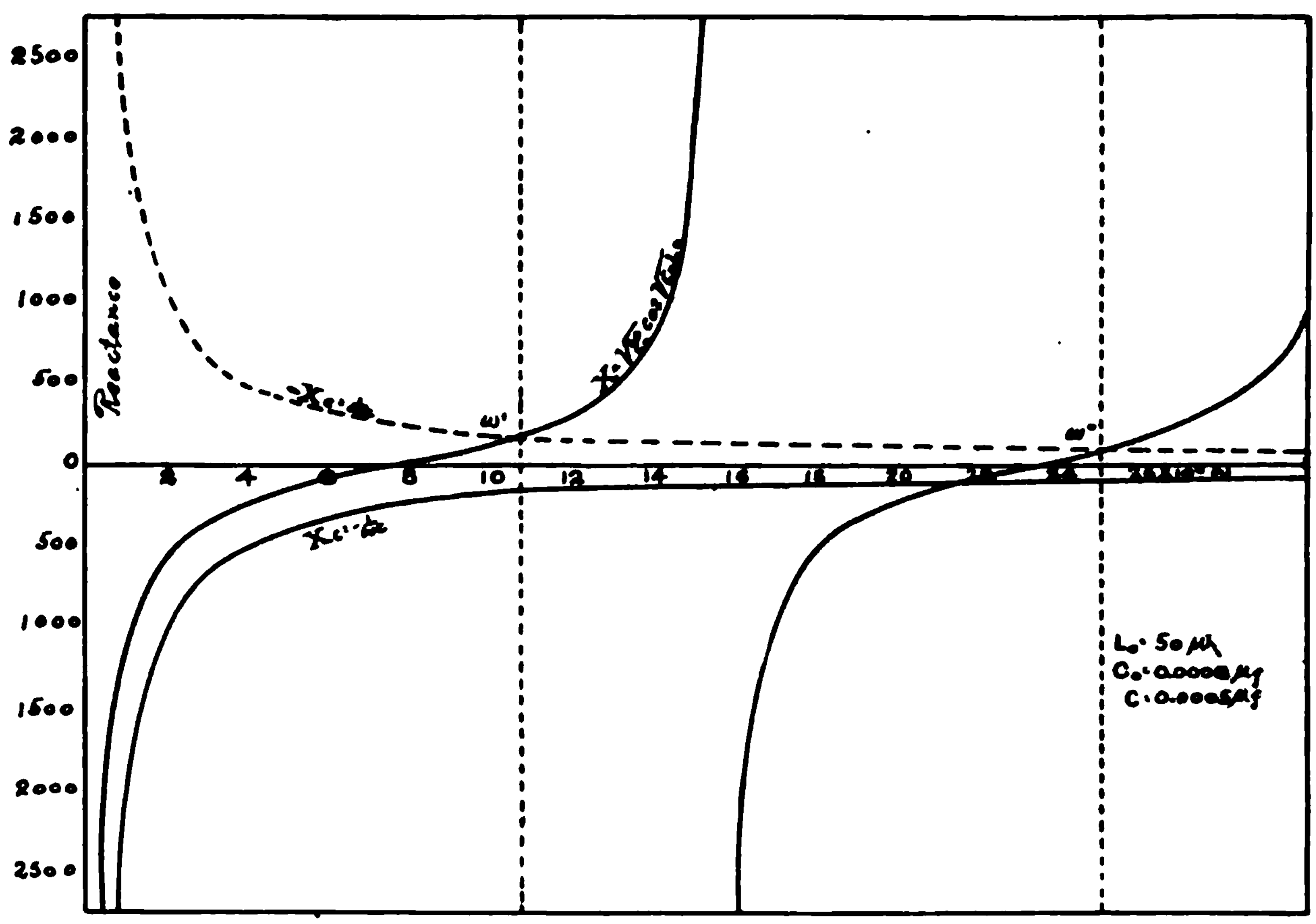

Figure 5—Curves of Aerial and Series Condenser Reactances

of the negative of this curve (drawn in dashed line) with the cotangent curves representing X gives the frequencies for which

[2] Guyau, A.; "Lumière Electrique," 15, p. 13; 1911.
[3] Cohen, L.; "Electrical World," 65, p. 286; 1915.

TABLE I

DATA FOR LOADED ANTENNA CALCULATIONS

$\frac{L}{L_o}$	$\omega\sqrt{C_o L_o}$	$\frac{1}{\sqrt{\frac{L}{L_o}+\frac{1}{3}}}$	Difference, per cent	$\frac{L}{L_o}$	$\omega\sqrt{C_o L_o}$	$\frac{1}{\sqrt{\frac{L}{L_o}+\frac{1}{3}}}$	Difference, per cent
0.0	1.571	1.732	10.3	3.1	0.539	0.540	0.1
.1	1.429	1.519	6.3	3.2	.532	.532	.1
.2	1.314	1.369	4.2	3.3	.524	.525	.1
.3	1.220	1.257	3.0	3.4	.517	.518	.1
.4	1.142	1.168	2.3	3.5	.510	.511	.1
.5	1.077	1.095	1.7	3.6	.504	.504	.0
.6	1.021	1.035	1.4	3.7	.4977	.4979	.0
.7	.973	.984	1.1	3.8	.4916	.4919	.0
.8	.931	.939	.9	3.9	.4859	.4860	.0
.9	.894	.900	.7	4.0	.4801	.4804	.0
1.0	.860	.866	.7	4.5	.4548	.4549	.0
1.1	.831	.835	.5	5.0	.4330	.4330	.0
1.2	.804	.808	.5	5.5	.4141	....	....
1.3	.779	.782	.4	6.0	.3974	....	....
1.4	.757	.760	.4	6.5	.3826	....	....
1.5	.736	.739	.4	7.0	.3693	....	....
1.6	.717	.719	.3	7.5	.3574	....	....
1.7	.699	.701	.3	8.0	.3465	....	....
1.8	.683	.685	.3	8.5	.3366	....	....
1.9	.668	.689	.3	9.0	.3275	....	....
2.0	.653	.655	.3	9.5	.3189	....	....
2.1	.640	.641	.2	10.0	.3111	....	....
2.2	.627	.628	.2	11.0	.2972	....	....
2.3	.615	.616	.2	12.0	.2850	....	....
2.4	.604	.605	.2	13.0	.2741	....	....
2.5	.593	.594	.2	14.0	.2644	....	....
2.6	.583	.584	.2	15.0	.2556	....	....
2.7	.574	.574	.2	16.0	.2476	....	....
2.8	.564	.565	.1	17.0	.2402	....	....
2.9	.556	.556	.1	18.0	.2338	....	....
3.0	.547	.548	.1	19.0	.2277	....	....
				20.0	.2219	....	....

$X_c + X = 0$, and hence the natural frequencies of oscillation of the antenna. The frequencies are increased (the wave length decreased) by the insertion of the condenser and the oscillations of higher frequencies are not integral multiples of the lowest.

The condition $X_c + X = 0$ is expressed by the equation

$$-\frac{\tan \omega \sqrt{C_o L_o}}{\omega \sqrt{C_o L_o}} = \frac{C}{C_o} \tag{9}$$

which has been given by Guyau. Equation (9) may be solved graphically as above or a table similar to Table I may be prepared giving $\omega \sqrt{C_o L_o}$ for different values of $\frac{C}{C_o}$. More complicated circuits may be solved in a similar manner.

3. Effective Resistance, Inductance, and Capacity

In the following, the most important practical case of a loading coil in the lead-in and the natural oscillation of lowest frequency will alone be considered. The problem is to replace the antenna of Figure 6 (a) which has a loading coil L in the lead-in and an aerial with distributed characteristics by a circuit Figure 6 (b) consisting of the inductance L in series with lumped resistance R_e, inductance L_e, capacity C_e, which are equivalent to the aerial. It is necessary, however, to state how these effective values are to be defined.

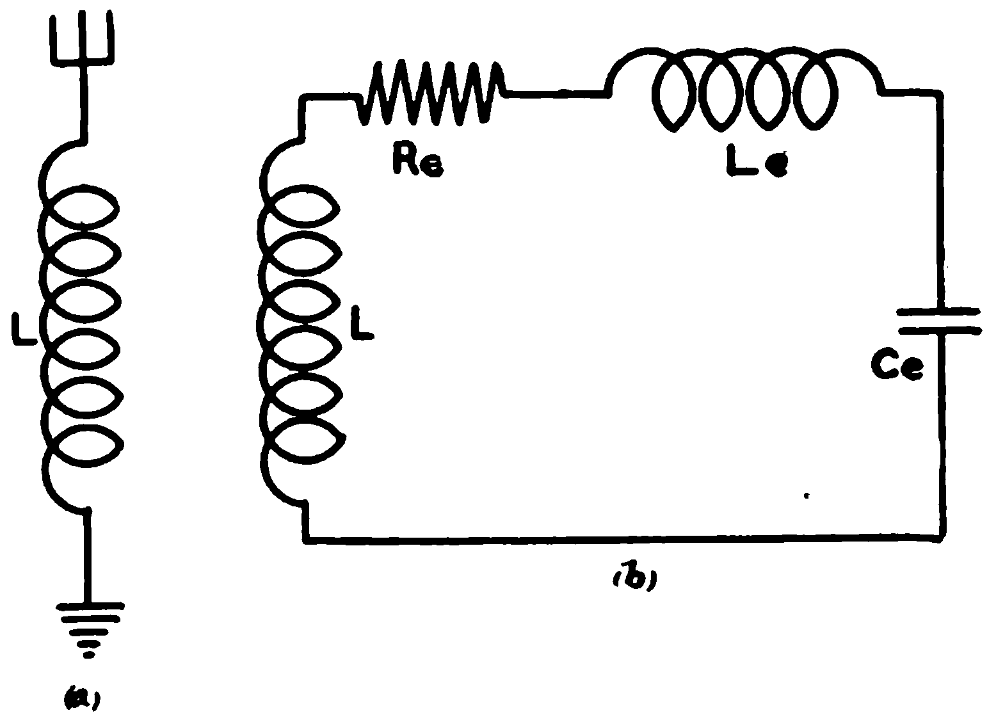

Figure 6

In practice the quantities which are of importance in an antenna are the resonant wave length or frequency and the current at the current maximum. The quantities L_e and C_e are, therefore, defined as those which will give the circuit (b) the

same resonant frequency as the antenna in (a). Further the three quantities L_e, C_e, and R_e must be such that the current in (b) will be the same as the maximum in the antenna for the same applied emf. whether undamped or damped with any decrement. These conditions determine L_e, C_e, and R_e uniquely at any given frequency, and are the proper values for an artificial antenna which is to represent an actual antenna at a particular frequency. In the two circuits the corresponding maxima of magnetic energies and electrostatic energies and the dissipation of energy will be the same.

Zenneck[4] has shown how these effective values of inductance capacity and resistance can be computed when the current and voltage distributions are known. Thus, if at any point x on the oscillator, the current i and the voltage v are given by

$$i=I\,f\,(x)\,;\,v=V\,\phi\,(x)$$

where I is the value of the current at the current loop and V the maximum voltage, then the differential equation of the oscillation is

$$\frac{\partial I}{\partial t}\int R_1 f\,(x)^2\,d\,x+\frac{\partial^2 I}{\partial t^2}\int L_1 f\,(x)^2\,d\,x+\frac{1}{\frac{\{\int C_1\phi(x)\,d\,x\}^2}{\int C_1\,\phi\,(x)^2\,d\,x}}=0$$

where the integrals are taken over the whole oscillator. If we write

$$R_e=\int R_1 f\,(x)^2\,d\,x \tag{10}$$

$$L_e=\int L_1 f\,(x)^2\,d\,x \tag{11}$$

$$C_e=\frac{\{\int C_1\,\phi\,(x)\,d\,x\}^2}{\int C_1\,\phi\,(x)^2\,d\,x} \tag{12}$$

the equation becomes

$$R_e\frac{\partial I}{\partial t}+L_e\frac{\partial^2 I}{\partial t^2}+\frac{I}{C_e}=0$$

which is the differential equation of oscillation of a simple circuit with lumped resistance, inductance, and capacity of values R_e, L_e, and C_e and in which the current is the same as the maximum in the distributed case. In order to evaluate these quantities, it is necessary only to determine $f\,(x)$ and $\phi\,(x)$; that is, the functions which specify the distribution of current and voltage on the oscillator. In this connection it will be assumed that the resistance is not of importance in determining these distributions.

[4] Zenneck, "Wireless Telegraphy" (Translated by A. E. Seelig), Note 40, p. 410.

At the far end of the aerial the current is zero, that is for $x=l$; $i_l=0$. From equations (3) and (4) for $x=l$

$$v_l=\cos\omega t\,(A\cos\omega\sqrt{C_1L_1}\,l+B\sin\omega\sqrt{C_1L_1}\,l)$$

$$i_l=\sin\omega t\sqrt{\frac{C_1}{L_1}}\,(A\sin\omega\sqrt{C_1L_1}\,l-B\cos\omega\sqrt{C_1L_1}\,l)$$

and since $i_l=0$

$$A\sin\omega\sqrt{C_1L_1}\,l=B\cos\omega\sqrt{C_1L_1}\,l$$

From (3) then we obtain

$$v=v_l\cos(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)$$

Hence $$\phi(x)=\cos(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)$$

Now for $x=0$ from (4) we obtain

$$i_o=-B\sqrt{\frac{C_1}{L_1}}\sin\omega t=-A\sqrt{\frac{C_1}{L_1}}\tan\omega\sqrt{C_1L_1}\,l\sin\omega t$$

whence $$i=i_o\frac{\sin(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)}{\sin\omega\sqrt{C_1L_1}\,l}$$

and $$f(x)=\frac{\sin(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)}{\sin\omega\sqrt{C_1L_1}\,l}$$

We can now evaluate the expressions (10), (11), and (12). From (10)

$$\begin{aligned}R_e&=\int_o^l R_1\frac{\sin^2(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)\,dx}{\sin^2\omega\sqrt{C_1L_1}\,l}\\&=\frac{R_1}{\sin^2\omega\sqrt{C_1L_1}\,l}\left[\frac{l}{2}-\frac{\sin 2\omega\sqrt{C_1L_1}\,l}{4\omega\sqrt{C_1L_1}}\right]\\&=\frac{R_o}{2}\left[\frac{1}{\sin^2\omega\sqrt{C_oL_o}}-\frac{\cot\omega\sqrt{C_oL_o}}{\omega\sqrt{C_oL_o}}\right]\end{aligned}\qquad(13)$$

and from (11) which contains the same form of integral

$$L_e=\frac{L_o}{2}\left[\frac{1}{\sin^2\omega\sqrt{C_oL_o}}-\frac{\cot\omega\sqrt{C_oL_o}}{\omega\sqrt{C_oL_o}}\right]\qquad(14)$$

and from (12)

$$\begin{aligned}C_e&=\frac{\{\int_o^l C_1\cos(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)\,dx\}^2}{\int_o^l C_1\cos^2(\omega\sqrt{C_1L_1}\,l-\omega\sqrt{C_1L_1}\,x)\,dx}\\&=\frac{C_1^2\dfrac{\sin^2\omega\sqrt{C_1L_1}\,l}{(\omega\sqrt{C_1L_1})^2}}{C_1\left(\dfrac{l}{2}+\dfrac{\sin 2\omega\sqrt{C_1L_1}\,l}{4\omega\sqrt{C_1L_1}}\right)}\\&=\frac{C_o}{\left[\dfrac{\omega\sqrt{C_oL_o}\cot\omega\sqrt{C_oL_o}}{2}+\dfrac{\omega^2L_oC_o}{2\sin^2\omega\sqrt{C_oL_o}}\right]}\end{aligned}\qquad(15)$$

The expressions (14) and (15) should lead to the same value for the reactance X of the aerial as obtained before. It is readily shown that

$$X = \omega L_e - \frac{1}{\omega C_e} = -\sqrt{\frac{L_o}{C_o}} \cot \omega \sqrt{C_o L_o}$$

agreeing with equation (6).

It is of interest to investigate the values of these quantities at very low frequencies ($\omega \doteq 0$), frequently called the static values, and those corresponding to the natural frequency of the unloaded antenna or the so-called fundamental of the antenna. Substituting $\omega = 0$ in (13), (14), and (15) and evaluating the indeterminant which enters in the first two cases we obtain for the low frequency values

$$R_e = \frac{R_o}{3}$$

$$L_e = \frac{L_o}{3}$$

$$C_e = C_o \qquad (16)$$

At low frequencies, the current is a maximum at the lead-in end of the aerial and falls off linearly to zero at the far end. The effective resistance and inductance are one-third of the values which would obtain if the current were the same thruout. The voltage is, however, the same at all points and hence the effective capacity is the capacity per unit length times the length or C_o.

At the fundamental of the antenna, the reactance X of equation (6) becomes equal to zero and hence $\omega \sqrt{C_o L_o} = \frac{\pi}{2}$. Substituting this value in (13), (14), and (15)

$$\left.\begin{aligned} R_e &= \frac{R_o}{2} \\ L_e &= \frac{L_o}{2} \\ C_e &= \frac{8}{\pi^2} C_o \end{aligned}\right\} \qquad (17)$$

Hence in going from low frequencies up to that of the fundamental of the antenna, the resistance (neglecting radiation and skin effect) and the inductance (neglecting skin effect) increase by fifty per cent., the capacity, however, decreases by about twenty per cent. The incorrect values $\frac{2}{\pi} L_o$ and $\frac{2}{\pi} C_o$ have been fre-

quently given and commonly used as the values of the effective inductance and capacity of the antenna at its fundamental. These lead also to the incorrect value $L_e = \frac{L_o}{2}$ for the low frequency inductance[5].

The values for other frequencies may be obtained by substitution in (13), (14), (15). If the value L of the loading coil in the lead-in is given, the quantity $\omega\sqrt{C_o L_o}$ is directly obtained from Table 1.

4. Equivalent Circuit with Lumped Constants

Insofar as the frequency or wave length is concerned, the aerial of the antenna may be considered to have constant values of inductance and capacity and the values of frequency or wave length for different loading coils may be computed with slight error using the simple formula applicable to circuits with lumped inductance and capacity. The values of inductance and capacity ascribed to the aerial are the static or low frequency, that is, $\frac{L_o}{3}$ for the inductance and C_o for the capacity. The total inductance in case the loading coil has a value L will be $L + \frac{L_o}{3}$ and the frequency is given by

$$f = \frac{1}{2\pi\sqrt{\left(L + \frac{L_o}{3}\right) C_o}} \tag{18}$$

or the wave length in meters by

$$\lambda = 1884\sqrt{\left(L + \frac{L_o}{3}\right) C_o} \tag{19}$$

where the inductance is expressed in microhenrys and the capacity in microfarads. The accuracy with which this formula gives the wave length can be determined by comparison with the exact formula (8). In the second column of Table I are given

[5] These values are given by J. H. Morecroft in "Proc. I. R. E." 5, p. 389, 1917. It may be shown that they lead to correct values for the reactance of the aerial and hence to correct values of frequency as was verified by the experiments. They are not, however, the values which would be correct for an artificial antenna in which the current must equal the maximum in the actual antenna and in which the energies must also be equal to those in the antenna. The resistance values given by Prof. Morecroft agree with these requirements and with the values obtained here.

Values for the effective inductance and capacity in agreement with those of equation (17) above have been given by G. W. O. Howe, "Yearbook of Wireless Telegraphy and Telephony," page 699, 1917.

the values of $\omega\sqrt{C_0 L_0}$ for different values of L_0 as computed by formula (8). Formula (18) may be written in the form

$$\omega\sqrt{L_0 C_0} = \frac{1}{\sqrt{\frac{L}{L_0} + \frac{1}{3}}}$$

so that the values of $\omega\sqrt{C_0 L_0}$, which are proportional to the frequency, may readily be computed from this formula also. These values are given in the third column of Table I and the per cent. difference in the fourth column. It is seen that formula (18) gives values for the frequency which are correct to less than a per cent., excepting when very close to the fundamental of the antenna, i. e., for very small values of L. Under these conditions the simple formula leads to values of the frequency which are too high. Hence to the degree of accuracy shown, which is amply sufficient in most practical cases, *the aerial can be represented by its static inductance $\frac{L_0}{3}$ with its static capacity C_0 in series, and the frequency of oscillation with a loading coil L in the lead-in can be computed by the ordinary formula applicable to circuits with lumped constants.*

In an article by L. Cohen,[6] which has been copied in several other publications, it was stated that the use of the simple wave length formula would lead to very large errors when applied to the antenna with distributed constants. The large errors found by Cohen are due to his having used the value L_0 for the inductance of the aerial, instead of $\frac{L_0}{3}$, in applying the simple formula.

IV. The Inductance Coil

The transmission line theory can also be applied to the treatment of the effects of distributed capacity in inductance coils. In Figure 7 (a) is represented a single layer solenoid connected to a variable condenser C. A and B are the terminals of the coil, D the middle, and the condensers drawn in dotted lines are supposed to represent the capacities between the different parts of the coil. In Figure 7 (b) the same coil is represented as a line with uniformly distributed inductance and capacity. These assumptions are admittedly rough, but are somewhat justified by the known similarity of the oscillations in long solenoids to those in a simple antenna.

[6] See foot-note 3.

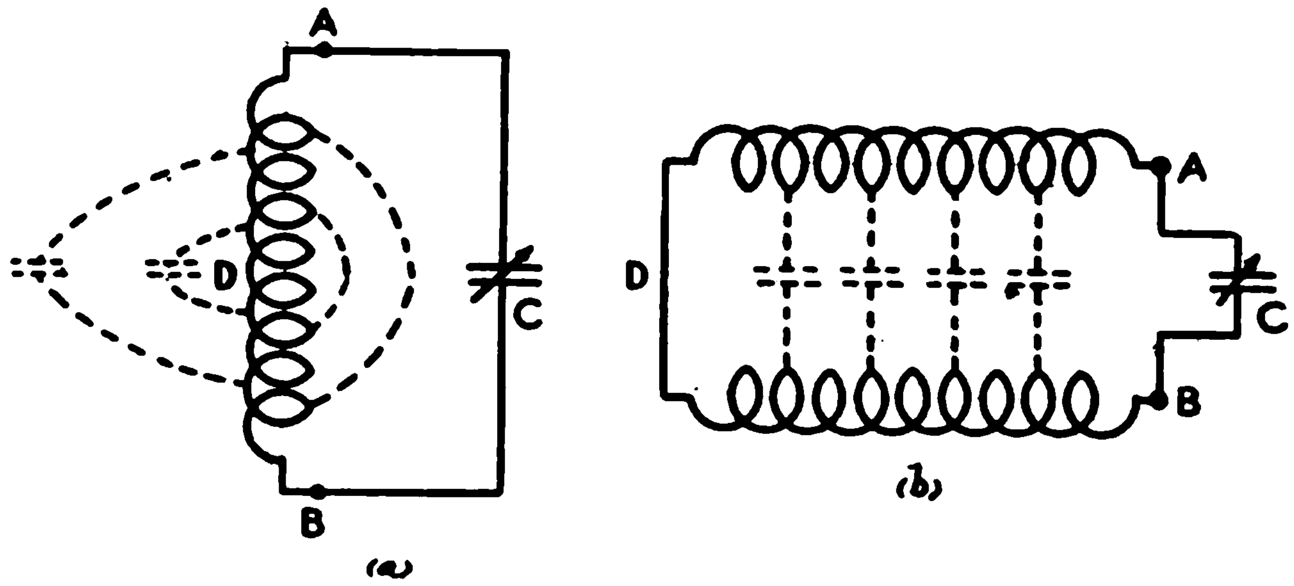

FIGURE 7—Inductance Coil Represented as a Line with Uniform Distribution of Inductance and Capacity

1. Reactance of the Coil

Using the same notation as before, an expression for the reactance of the coil, regarded from the terminals $A\,B\,(x=0)$, will be determined considering the line as closed at the far end $D\,(x=l)$. Equations (3) and (4) will again be applied, taking account of the new terminal condition, that is, for $x=l$; $v=0$. Hence

$$A \cos \omega \sqrt{C_1 L_1}\, l = -B \sin \omega \sqrt{C_1 L_1}\, l$$

and for $x=0$

$$v_o = A \cos \omega t = -B \tan \omega \sqrt{C_1 L_1}\, l \cos \omega t$$

$$i_o = -\sqrt{\frac{C_1}{L_1}} B \sin \omega t$$

which gives for the reactance of the coil regarded from the terminals $A\,B$,

$$X' = \sqrt{\frac{L_1}{C_1}} \tan \omega \sqrt{C_1 L_1}\, l$$

or

$$X' = \sqrt{\frac{L_o}{C_o}} \tan \omega \sqrt{C_o L_o} \qquad (20)$$

2. Natural Frequencies of Oscillation

At low frequencies, the reactance of the coil is very small and positive, but increases with increasing frequency and becomes infinite when $\omega \sqrt{C_o L_o} = \frac{\pi}{2}$. This represents the lowest frequency of natural oscillation of the coil when the terminals are open. Above this frequency the reactance is highly negative, approaching zero at the frequency $\omega \sqrt{C_o L_o} = \pi$. In this range of frequencies, the coil behaves as a condenser and would require

an inductance across the terminals to form a resonant circuit. At the frequency $\omega\sqrt{C_oL_o}=\pi$ the coil will oscillate with its terminals short-circuited. As the frequency is still further increased the reactance again becomes increasingly positive.

(a) Condenser Across the Terminals. The natural frequencies of oscillation of the coil when connected to a condenser C are given by the condition that the total reactance of the circuit shall be zero.

$$X'+X_c=0$$

From this we have

$$\sqrt{\frac{L_o}{C_o}}\tan\omega\sqrt{C_oL_o}=\frac{1}{\omega C}$$

or

$$\frac{\cot\omega\sqrt{C_oL_o}}{\omega\sqrt{C_oL_o}}=\frac{C}{C_o} \tag{21}$$

This expression is the same as (8) obtained in the case of the loaded antenna, excepting that $\frac{C}{C_o}$ occurs on the right-hand side instead of $\frac{L}{L_o}$, and shows that the frequency is decreased and wave length increased by increasing the capacity across the coil in a manner entirely similar to the decrease in frequency produced by inserting loading coils in the antenna lead-in.

3. Equivalent Circuit with Lumped Constants

It is of interest to investigate the effective values of inductance and capacity of the coil at very low frequencies. Expanding the tangent in equation (20) into a series we find

$$X'=\omega L_o\left(1+\frac{\omega^2C_oL_o}{3}+\ .\ .\ .\ .\right)$$

and neglecting higher power terms this may be written

$$X'=\frac{(\omega L_o)\left(-\frac{3}{\omega C_o}\right)}{\omega L_o-\frac{3}{\omega C_o}}$$

This is the reactance of an inductance L_o in parallel with a capacity $\frac{C_o}{3}$ which shows that at low frequencies the coil may be regarded as an inductance L_o with a capacity $\frac{C_o}{3}$ across the terminals and, therefore, in parallel with the external condenser

C. Since at low frequencies the current is uniform thruout the whole coil, it is self evident that its inductance should be L_o.

Now the similarity between equations (21) and (8) shows that, just as accurately as in the similar case of the loaded antenna, the frequency of oscillation of a coil with *any capacity* C across the terminals is given by the formula

$$f=\frac{1}{2\pi\sqrt{L_o\left(C+\frac{C_o}{3}\right)}} \tag{22}$$

This, however, is *also* the expression for the frequency of a coil of pure inductance L_o with a capacity $\frac{C_o}{3}$ across its terminals and which is in parallel with an external capacity C. Therefore, insofar as frequency relations are concerned, *an inductance coil with distributed capacity is closely equivalent at any frequency to a pure inductance, equal to the low frequency inductance (neglecting skin effect), with a constant capacity across its terminals.* This is a well-known result of experiment[7] at least in the case of single layer solenoids which, considering the changes in current and voltage distribution in the coil with changing frequency, is not otherwise self-evident.

V. Antenna Measurements

1. Determination of Static Capacity and Inductance

In applying formula (8) to calculate the frequency of a loaded antenna, a knowledge of the quantities L_o and C_o is required. In applying formula (18), $\frac{L_o}{3}$ and C_o are required. Hence either formula may be used if the static capacity and inductance values are known. We will call these values simply the capacity C_a and inductance L_a of the antenna. Hence $C_a=C_o$, $L_a=\frac{L_o}{3}$ and the wave length from (19) is given by

$$\lambda=1884\sqrt{(L+L_a)\,C_a} \tag{23}$$

where inductance is expressed in microhenrys and capacity in microfarads as before.

The capacity and inductance of the antenna are then readily determined experimentally by the familiar method of inserting,

[7] G. W. O. Howe; "Proc. Phys. Soc.," London, 24, p. 251, 1912.
F. A. Kolster; "Proc. Inst. Radio Engrs.," 1, p. 19, 1913.
J. C. Hubbard; "Phys. Rev., 9, p. 529," 1917.

one after the other, two loading coils of known values L_1 and L_2 in the lead-in, and determining the frequency of oscillation or wave length for each. From the observed wave lengths λ_1 and λ_2 and known values of the inserted inductances, the inductance of the antenna is given by

$$L_a = \frac{L_1 \lambda_2^2 - L_2 \lambda_1^2}{\lambda_1^2 - \lambda_2^2} \tag{24}$$

and the capacity of the antenna from either

$$\left.\begin{aligned} \lambda_1 &= 1884 \sqrt{(L_1+L_a)\, C_a} \\ \lambda_2 &= 1884 \sqrt{(L_2+L_a)\, C_a} \end{aligned}\right\} \tag{25}$$

using preferably, the equation corresponding to the larger valued coil. This assumes that formula (23) holds exactly.

As an example let us assume that the antenna has $L_o = 50$ microhenrys and $C_o = 0.001$ microfarad, and that we insert two coils of 50 and 150 microhenrys and determine the wave lengths experimentally. We know from formula (8) and Table I that the wave lengths would be found to be 491 and 771 meters. From the observed wave lengths and known inductances, the value of L_a would be found by (24) to be

$$L_a = 17.8 \text{ microhenrys}$$

and from (25)

$$C_a = 0.00099_9 \text{ microfarad.}$$

C_a is very close to the assumed value of C_o but L_a differs by seven per cent. from $\frac{L_o}{3}$. This accuracy would ordinarily be sufficient. We can, however, by a second approximation, derive from the experimental data a more accurate value of L_a. For, the observed value of L_a furnishes rough values of $\frac{L_1}{L_o}$ and $\frac{L_2}{L_o}$ which in this example come out 0.96 and 2.88, respectively. But Table I gives the per cent. error of formula (23) for different values of $\frac{L}{L_o}$ and shows that this formula gives a 0.7 per cent. shorter wave length than 491 meters (or 488 meters) for $\frac{L}{L_o} = 0.96$ but no appreciable difference for $\frac{L}{L_o} = 2.88$. Recomputing L_a using 488 and 771 meters gives

$$L_a = 0.0168,$$

which is practically identical with the assumed $\frac{L_o}{3}$.

2. Determination of Effective Resistance, Inductance, and Capacity

When a source of undamped oscillations in a primary circuit induces current in a secondary tuned circuit, the current in the secondary, for a given emf. depends only upon the resistance of the secondary circuit. When damped oscillations are supplied by the source in the primary, the current in the secondary, for a given emf. and primary decrement, depends upon the decrement of the secondary, i. e., upon the resistance and ratio of capacity to inductance. The higher the decrement of the primary circuit relative to the decrement of the secondary, the more strongly does the current in the secondary depend upon its own decrement. This is evident from the expression for the current I in the secondary circuit

$$I^2 = \frac{NE_o^2}{4fR^2\delta'\left(1+\frac{\delta'}{\delta}\right)}$$

where δ' is the decrement of the primary, δ that of the secondary, R the resistance of the secondary, f the frequency, E_o the maximum value of the emf. impressed on the secondary, and N the wave train frequency.

These facts suggest a method of determining the effective resistance, inductance, and capacity of an antenna at a given frequency in which all of the measurements are made at one frequency, and which does not require any alteration of the antenna circuit whatsoever. The experimental circuits are arranged as shown in Figure 8, where S represents a coil in the primary circuit which may be thrown either into the circuit of

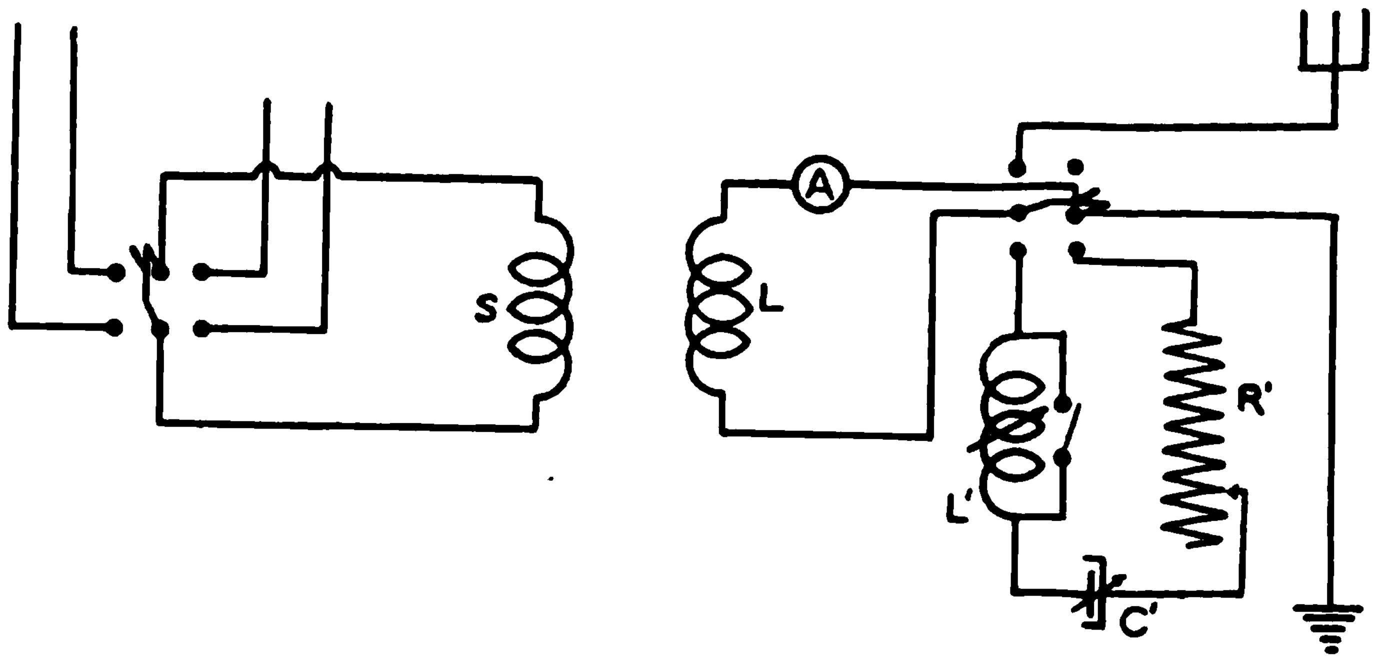

Figure 8—Circuits for Determining the Effective Resistance, Inductance, and Capacity of an Antenna

a source of undamped or of damped oscillations. The coil L is the loading coil of the antenna which may be thrown over to the measuring circuit containing a variable inductance L', a variable condenser C', and variable resistance R'. The condenser C' should be resistance-free and shielded, the shielded terminal being connected to the ground side. First the undamped source is tuned to the antenna, and then the $L'C'$ circuit tuned to the source. The resistance R' is then varied until the current is the same in the two positions. The resistance of the $L'C'$ circuit is then equal to R_e, the effective resistance of the aerial-ground portion of the antenna and $L'C' = L_eC_e$. Next the damped source is tuned to the antenna and the change in current noted when the connection is thrown over to the $L'C'$ circuit. If the current increases, the value of C' is greater than C_e, and vice versa. By varying both L' and C', keeping the tuning and R' unchanged, the current may be adjusted to the same value in both positions. Then since $L'C' = L_eC_e$ and $\frac{C'}{L'} = \frac{C_e}{L_e}$, the value of C' gives C_e and that of L' gives L_e. Large changes in the variometer setting may result in appreciable changes in its resistance, so that the measurement should be repeated after the approximate values have been found. To eliminate the resistance of the variometer in determining R_e, the variometer is short-circuited and, using undamped oscillations, the resonance current is adjusted to equality in the two positions by varying R'. Then $R' = R_e$. The measurement requires steady sources of feebly damped and strongly damped current. The former is readily obtained by using a vacuum tube generator. A resonance transformer and magnesium spark gap operating at a low spark frequency serves very satisfactorily for the latter source or a single source of which the damping can be varied will suffice. An accuracy of one per cent. is not difficult to obtain.

3. The Effect of Imperfect Dielectrics Upon the Resistance of an Antenna

The typical curve of the variation of the resistance of an antenna with the wave length of the oscillation is shown in Figure 10 (b). It has two characteristic features, a rapid decrease in resistance with increasing wave length in the region of the shorter waves and an apparent linear increase in resistance with increasing wave length at long waves. The decrease in resistance at short waves is ascribed mainly to the decrease in the power

radiated in the form of electro-magnetic waves. This so-called radiation resistance should by theory decrease in inverse ratio with the square of the wave length. Skin effect and the change in distribution of current along the antenna would likewise produce resistance variations such that the resistance would decrease with increasing wave length. It has been difficult, however, to account for the observed linear increase in resistance at the longer waves. Austin[8] pointed out the similarity in the linear increase in resistance of an antenna at long wave lengths with the behavior of an absorbing condenser and concluded that dielectric absorption was a probable explanation of the phenomenon.[9]

The fact that in the curves which he had obtained for ship stations, the rise in resistance was less marked than for land stations, led him to believe that the absorption was probably caused by the ground acting as an imperfect dielectric. Austin stated that if we consider the ground as a dielectric rather than a conductor and consider it as a portion of the total dielectric lying between the antenna regarded as the upper plate of a condenser, and a ground water regarded as a lower plate, we reach a very probable explanation of many antenna resistance curves. The measurements carried out by the author verify Austin's hypothesis that the effect is caused by dielectric absorption; but do not confirm the supposition that the absorbing dielectric in question is the ground. Figure 9 shows the values of the equivalent resistance obtained at telephone frequencies for a small flat-top antenna at the Bureau of Standards. This antenna runs from a building to a tree and has a capacity of 650 micro-microfarads (0.00065 microfarad). The measurements were made at wave lengths varying from 100,000 to 750,000 meters, the equivalent resistance increasing linearly from 1,000 to 9,000 ohms. This is the order of magnitude which would be expected from Austin's measurements at radio frequencies upon an antenna for which the rise in resistance was particularly marked.

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[8] L. W. Austin; "Bulletin, Bureau of Standards," 12, p. 465, 1915. "Jahrbuch d. drahtl. Tel.," 9, p. 498, 1915.

[9] In a perfect condenser, or one which shows no energy loss, the phase of the current I in 90° in advance of the electromotive force E. In an imperfect condenser, the power loss, however caused, is given by $IE \sin \theta$ where θ is the phase difference or the angle by which the current lags from quadrature. An equivalent power loss is occasioned by a resistance (ρ) in series with a perfect condenser when this equivalent resistance satisfies the relation $\tan \theta = C\, p\, \rho$ where C is the capacity and $p = 2\pi$ times the frequency. It is characteristic of a condenser with an absorbing dielectric that the phase difference θ is, roughly, independent of the frequency, and hence the equivalent resistance must vary inversely as the frequency or directly as the wave length.

It seemed impossible, however, to ascribe this absorption to the ground acting as an imperfect dielectric. As shown by the above data, the effect persists at telephone frequencies while the calculations of True [10] and Reich,[11] based upon the measurements of conductivity and dielectric constant of the ground as given by Zenneck,[12] show that even for so high a frequency as would correspond to a wave length of a 1,000 meters, the magnitude of the conduction current in the ground exceeds by a hundred times that of the displacement current. Further, the

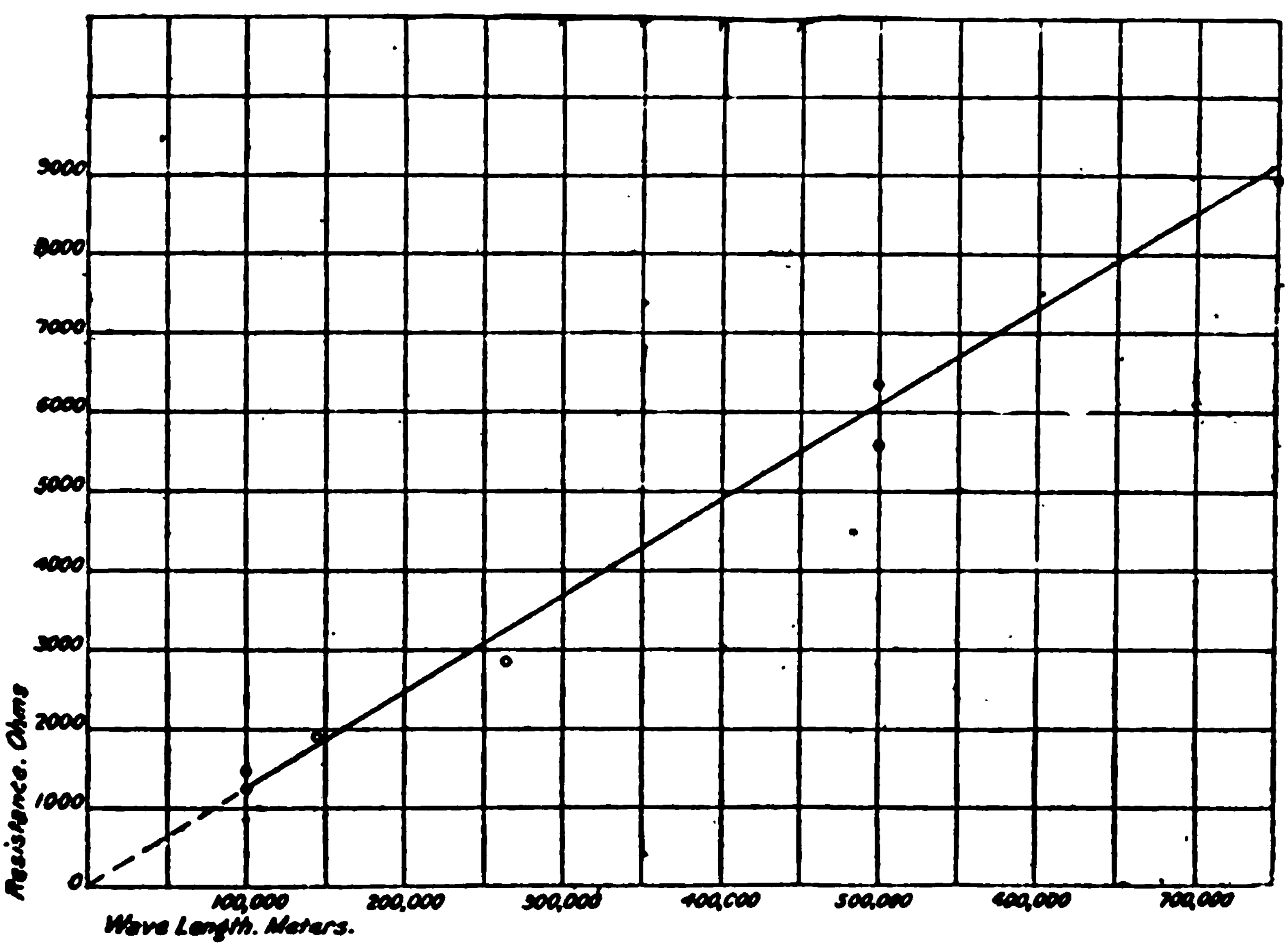

FIGURE 9—Equivalent Resistance of an Antenna at Telephone Frequencies

absence of absorption in the ground was also shown by measurements at telephone frequencies upon a guard plate condenser with part air and part clay between the plates. This condenser behaved as a perfect condenser with a series resistance that was independent of the frequency. Only when the clay was exceedingly dry and particularly when it was loosely packed, was there any indication of absorption.

The observed large effect upon the absorption of variable air condensers brought about by the poor dielectric properties

[10] H. True; "Jahrb. d. drahtl. Tel.," 5, p. 125; 1911-12.
[11] M. Reich; "Jahrb. d. drahtl. Tel.," 5, pp. 176, 253; 1911-12.
[12] J. Zenneck; "Ann. d. Phys.," 23, p. 859; 1907.

of small amounts of insulators in the electric field suggested to the author that the absorption in antennas is likewise caused by the presence of poor dielectrics in the field of the antenna. Accordingly, an experimental antenna was built in which the bad effects of poor dielectrics in the neighborhood were carefully avoided, but in which any absorption that might be caused by the ground would be considerably magnified. The main capacity of the antenna consisted of six parallel wires at a distance of about 0.3 of a meter (1 foot) above the surface of the ground and located at a considerable distance from the nearest building or tree. The antenna was supported by four wooden posts, but was insulated from them by double porcelain insulators spaced about a meter (3 feet) apart. A single lead, similarly insulated, ran to the building in which the measurements were made. The earth connection was made to the water pipes of the building. The proximity of the antenna wires to the ground should reduce the lateral spread of the electrostatic field and hence the displacement thru the wooden posts or other poor dielectrics, while the amount of ground between the antenna wires and ground water should be proportionately increased. The double-spaced insulators also served to reduce the capacity thru the supports. The capacity of the antenna was 850 micro-microfarads. The resulting resistance curve, for measurements made just within the window of the building, is shown in curve A of Figure 10. The rise in resistance even at 12,000 meters is very small, and probably caused by the lead wire to the building. The result was also verified at telephone frequencies where the absorption was barely detectable (less than 60 ohms at 3,000 cycles).

Curve B of the same figure shows the effect produced by adding a small capacity thru the wooden supports. Wires were run from the insulated portion of the antenna to porcelain insulators on three of the stakes, the total capacity being increased by only 40 micro-microfarads or less than 5 per cent. The effect of adding this small imperfect condenser is very marked. The linear increase in resistance becomes pronounced, and *brings with it an increase in the resistance of the antenna at all wave lengths.*

The effect of running the lead wires to an antenna inside of a building was also investigated. Curve C of Figure 11 shows the results upon the above described antenna under the same conditions as those obtaining for curve A of Figure 10 (reproduced in dash lines in Figure 11), excepting that the measurements were made within the room at a distance of about 5 meters

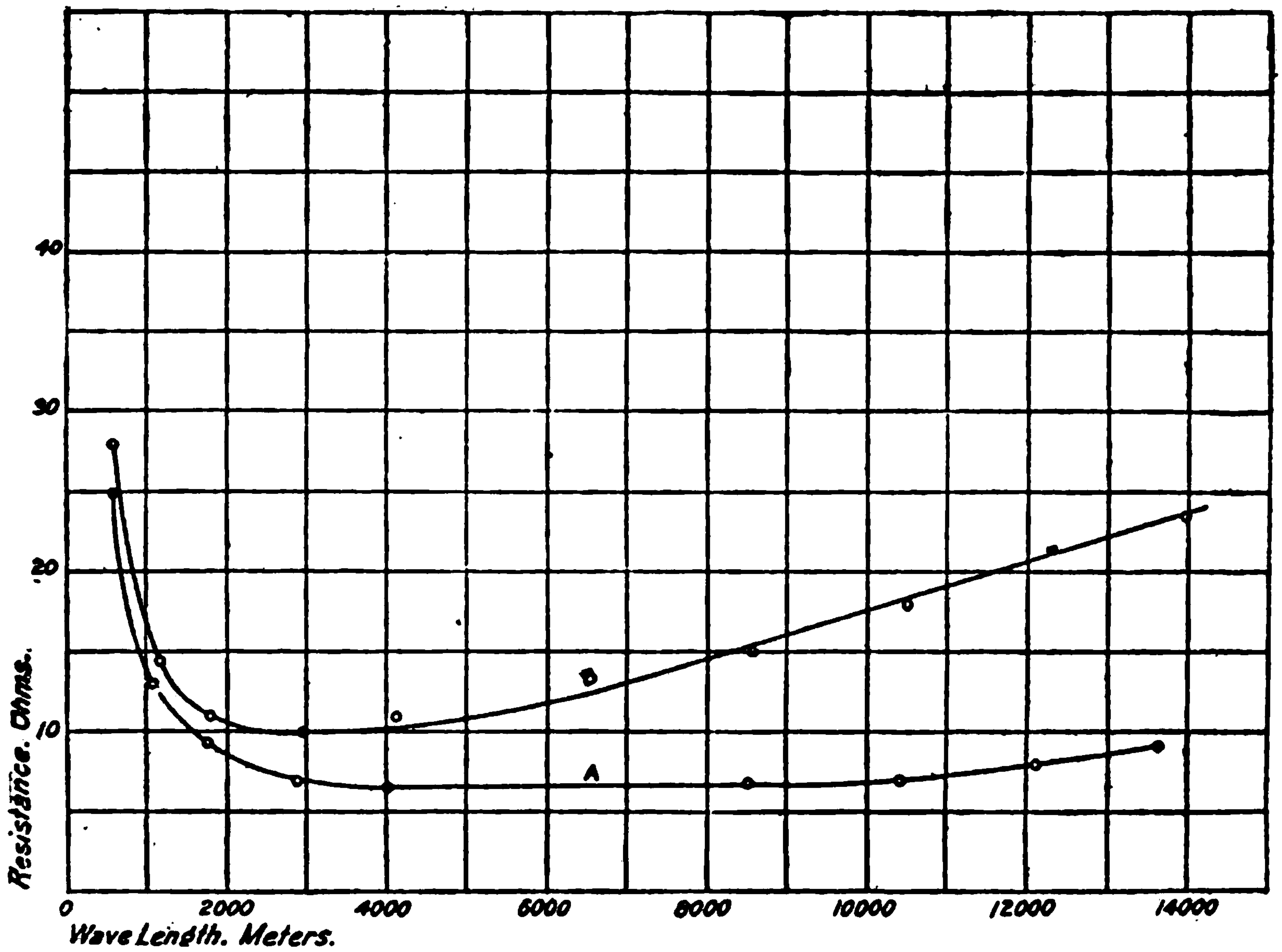

FIGURE 10—A, Resistance Curve for Antenna with Extremely Small Absorption; B, Effect of Adding Small Imperfect Capacity Through the Wooden Supports

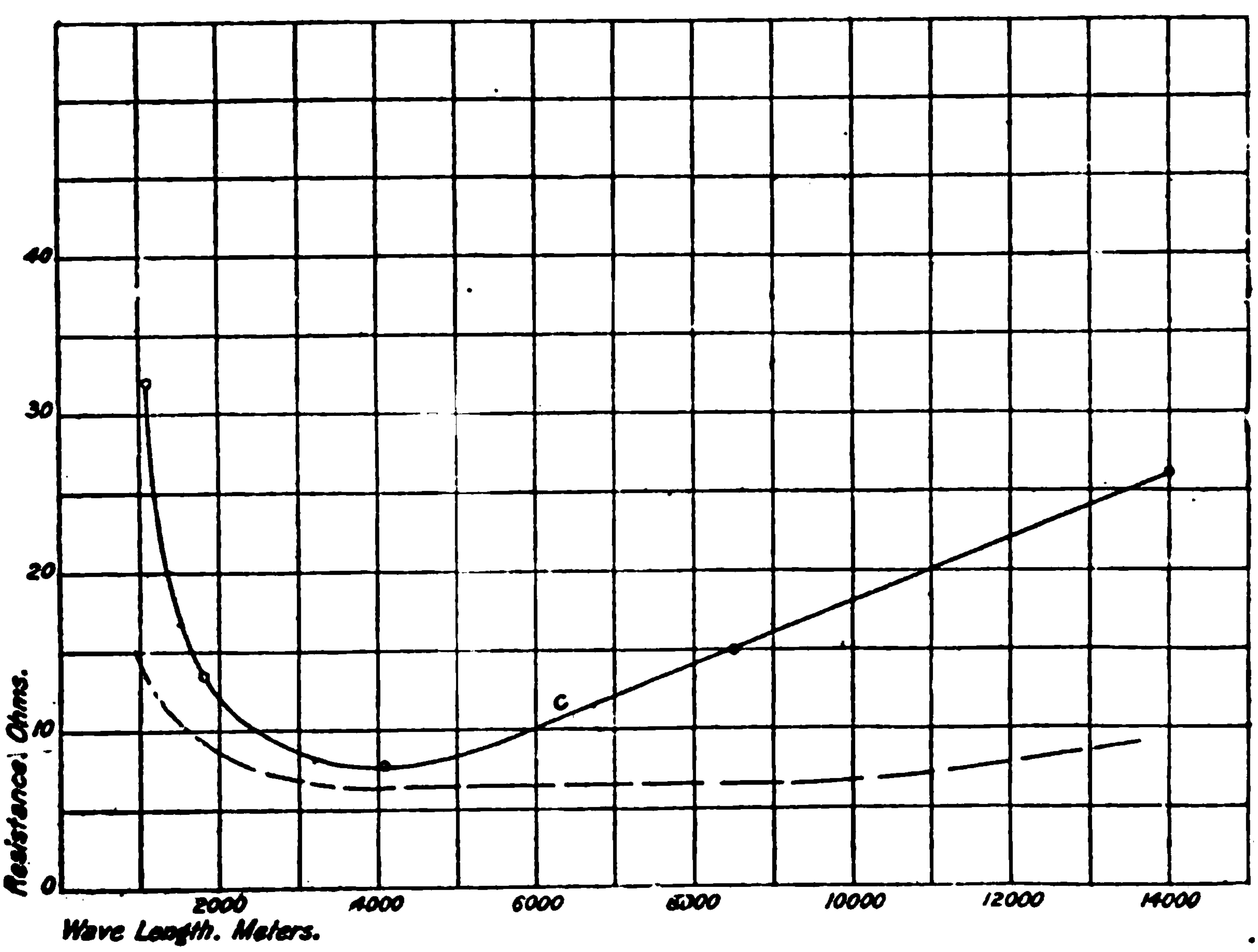

FIGURE 11—C, Absorption Effect Produced by the Portion of the Leads to an Antenna Inside of a Building

(16 feet) from the window. The increase in capacity in this case was 60 micro-microfarads. Curves D and E of Figure 12 were obtained for antennas completely within the building, the former having a capacity of 290 micro-microfarads, the latter having double the capacity. In this figure, the scale of resistance has been doubled. It is of interest to note that the equiva-

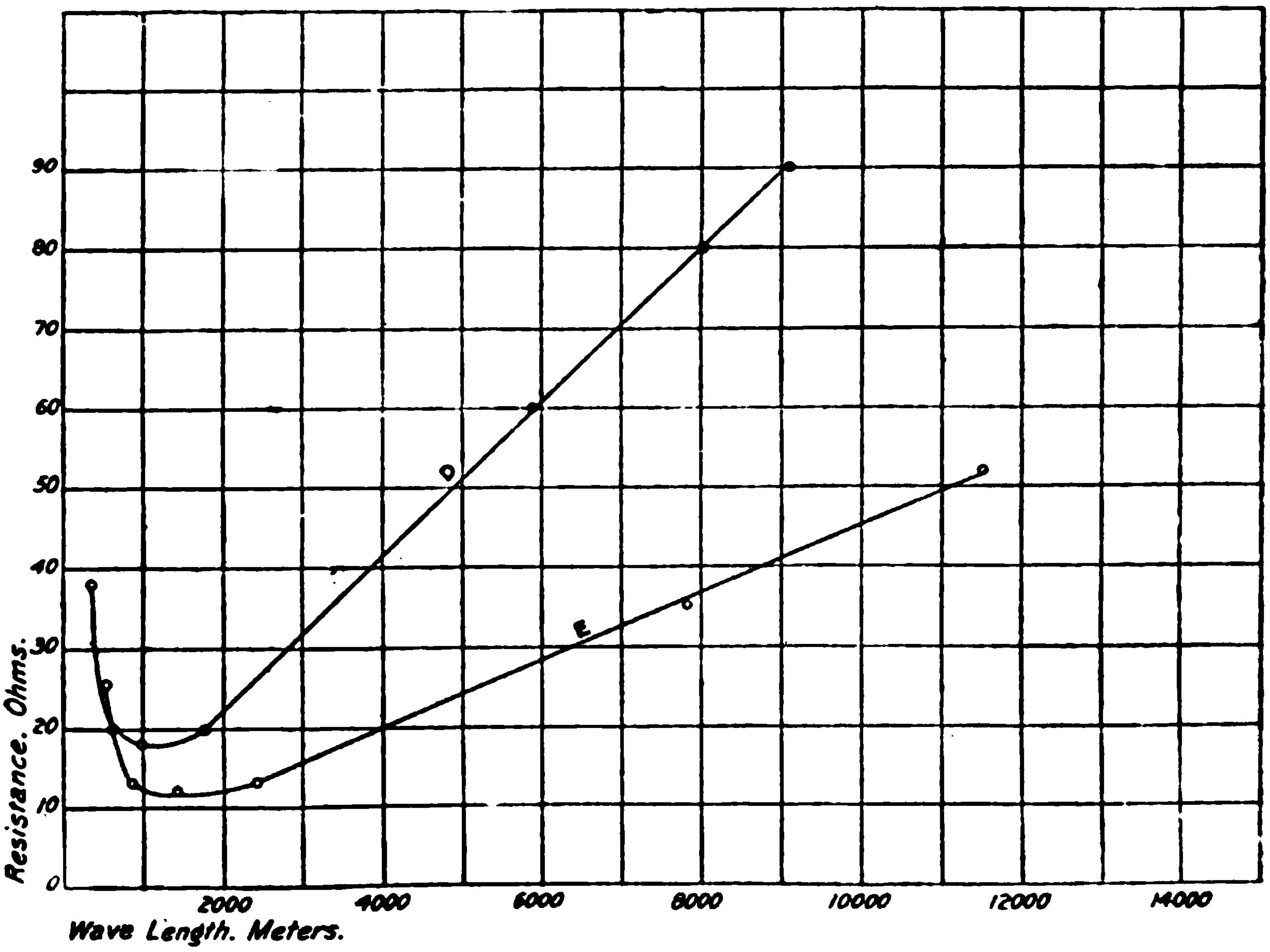

FIGURE 12—Resistance Curves for Antennas Completely Within a Building. The Capacity for E is Twice That for D

lent series resistance in the case of the smaller capacity is approximately double that for the capacity of twice the size, which is the requisite condition for absorbing condensers with the same phase difference. Measurements, at telephone frequencies, were also made upon an antenna consisting of three wires stretched vertically along the outside of a brick building at a distance of 0.4 of a meter (1 foot) from the wall. The phase difference of the condenser was about 15 minutes, corresponding to an equivalent resistance of about 35 ohms at 10,000 meters for the capacity of 800 micro-microfarads. The phase difference is about the same as that obtained in the case of antennas within the building.

Finally the effect of a tree upon the absorption of an antenna was investigated. An antenna consisting of two parallel wires

was strung from a building to a tree 20 meters (66 feet) distant at an average height of about 5 meters (16 feet) from the ground. The antenna terminated in a section of about 6 meters (20 feet) of wire which ran from limb to limb of the tree, but was insulated from it by porcelain insulators. At a distance of about 2 meters (6 feet) from the tree double porcelain insulators were interposed in each antenna wire, so that measurements could be made with the section in the tree included or excluded. Curve F of Figure 13 was obtained for the latter case, while curve G shows the enormous absorption produced by including the portion of the antenna in the tree. The capacity was increased

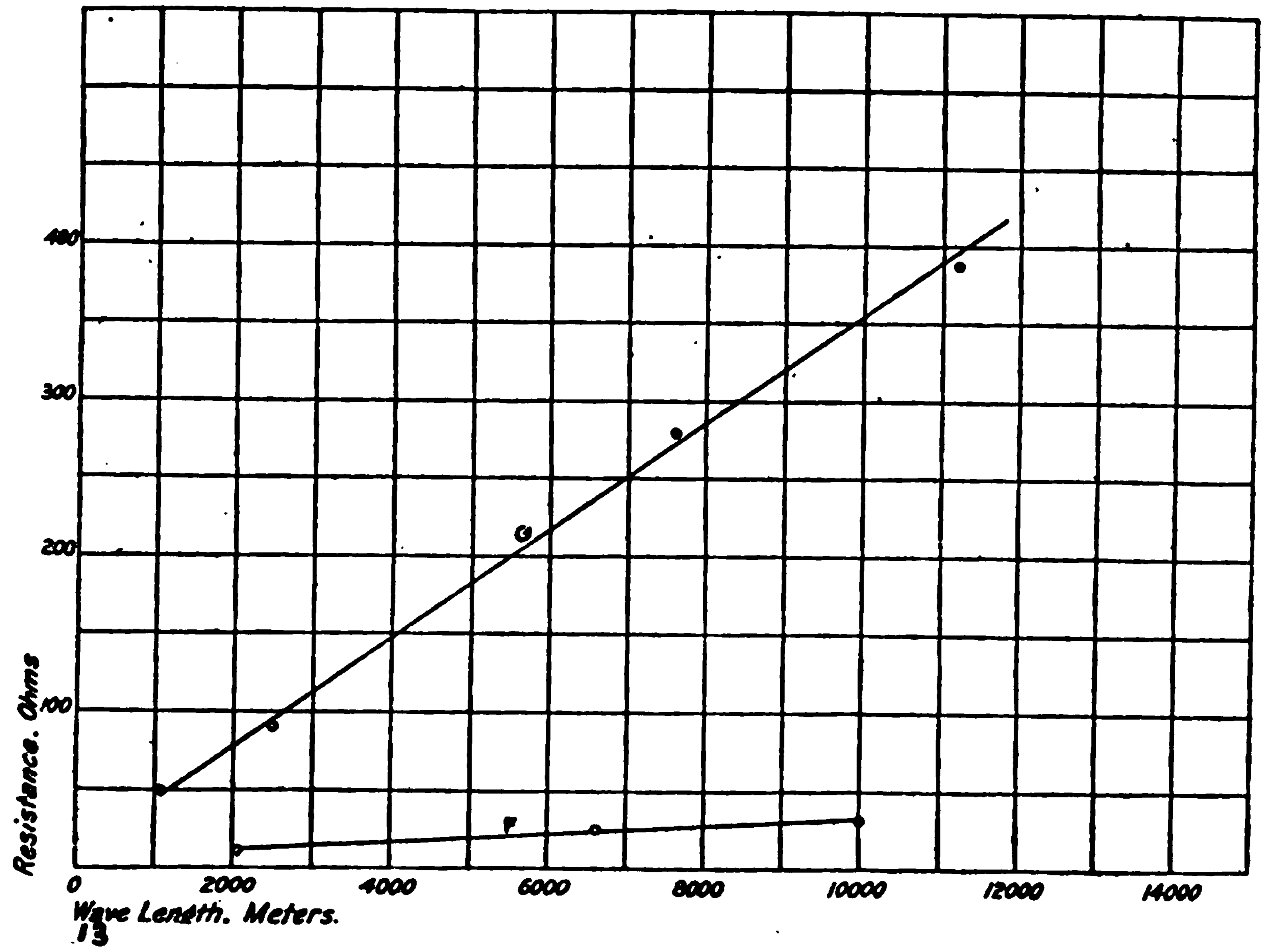

FIGURE 13—Curve G Shows the Enormous Absorption Caused by a Tree. Curve F Is the Same Antenna with Portion in Tree Excluded

from 390 to 540 micro-microfarads, and the phase difference of the condenser with the portion of the tree included was roughly 2 degrees. The measurements were made in the winter when the tree was free from foliage.

From the above it is evident that in the design of an antenna it is a matter of importance to keep the dielectric absorption of the antenna, regarded as a condenser, as low as possible in order to minimize the waste of energy in the antenna and so improve

its efficiency as a radiator. There is a possibility of greatly improving upon the design of existing antennas in this respect. The requirement is that the capacity thru wooden masts, trees, buildings, insulators, and so on, must be extremely small in comparison to the capacity of the antenna thru unobstructed air.

SUMMARY: After considering the theory of circuits having uniformly distributed constants, the author shows graphically the frequency-variation of reactance of such circuits and, after further analysis, those of inductance-loaded and capacity-loaded antennas as well.

The calculation of the effective constants of the antenna at radio frequencies in terms of their corresponding values at audio frequencies follows. Very simple relations are found, differing from those frequently given in the radio literature. The equivalent circuits of loaded and unloaded antennas are given, together with practical measuring methods for determining the effective constants at radio frequencies.

The frequency-variation of effective resistance of an antenna is then considered. Measurements are described and curves given which indicate that antenna resistance is largely due to imperfect dielectrics in the field of the antenna, and emphasis is placed on the necessity of avoiding such dielectrics in regions of strong antenna field.

FURTHER DISCUSSION ON "ON THE ELECTRICAL OPERATION AND MECHANICAL DESIGN OF AN IMPULSE EXCITATION MULTI-SPARK GROUP RADIO TRANSMITTER" BY BOWDEN WASHINGTON

By

Samuel Cohen

The paper by Mr. Bowden Washington in the December, 1918, Proceedings of The Institute of Radio Engineers contained some very interesting information regarding impact excitation transmitter and especially gaps that gave most suitable results for this kind of work.

The writer has conducted a number of experiments on impact excitation transmitter, and the data obtained verifies to a great extent the results of Ensign Washington. In reference to the tungsten electrode gap, the writer found that altho the gap functioned most regularly with a stiffer circuit, yet the tone emitted by the same was very poor. The tone corresponded very much to that of escaping whistling steam. This was obtained when operating the gap in open air. The tone effect was considerably improved by immersing the electrodes in alcohol and having the discharge take place therein. This improved considerably the general operating characteristics of the gap and it was possible to run it at much lower potential. Copper and tungsten electrodes in alcohol showed favorable results.

A type of gap which gave most satisfactory results for impact excitation work is a combination of copper and amalgamated copper. The results obtained were far better than the ones obtained with the use of the tungsten-tungsten and tungsten-copper electrodes. The type of gap operating in alcohol was compared with the Chaffee gap of copper aluminum electrodes under identical condition, and it was found that the copper-amalgamated copper electrodes proved much better both in efficiency and in note effect. It was also found that it can be used with a much stiffer circuit without affecting its operating qualities. The potential across the gap was somewhat lower than the Chaffee gap.

The only present disadvantage that the copper-amalgamated copper gap possesses is that the mercury on one of the electrodes becomes carbonized from the alcohol and thereby changing the operating characteristics. It takes about six to ten hours of continuous use before a sufficient carbonization takes place on the electrode to change the operating characteristics of the gap. The spacing of the electrodes was in the order of 0.013 inch (0.032 cm). The gap discharging surface was 3 inches (7.62 cm.) in diameter.

The tungsten-tungsten electrode gap immersed in alcohol seems to give excellent results for radio telephony. Some of the results thus far obtained were very satisfactory, and further experiments are to be conducted in this direction. The gap is very constant and regular in its operation and it works very satisfactorily on potentials of 100 volts direct current. The use of alcohol as the discharge medium is important, as the gap functions very poorly without it and requires a much higher operating potential.

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PROCEEDINGS

of

The Institute of Radio Engineers

(INCORPORATED)

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EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

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THE COLLEGE OF THE CITY OF NEW YORK

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The Institute of Radio Engineers

announces with regret the death of

Harold C. Schreiner

Mr. Schreiner was born February 9, 1895, in Chicago, Illinois. In June, 1917, he left the college which he was then attending in order to join the Signal Corps. At that time, he had completed three years' work of the course in electrical engineering. He had previously enlisted in the Signal Corps three days before the declaration of war by the United States.

In October, 1917, he was sent to Camp Custer, Battle Creek, Michigan. A month later, he was transferred to the Radio School at College Park, Maryland. His high grade in his work there led to his being commissioned as an officer. He was assigned to the 8th Signal Battalion, 4th Division, which was then stationed at Camp Green, Charlotte, North Carolina.

The battalion to which he was attached was ordered to France in May, 1918. He fought at Chateau Thierry, and later in the Argonne Sector. On September 26th, Lieutenant Schreiner was wounded. Within three weeks of this time, he succumbed to the effects of this injury. He was known among his associates and men as an officer of efficient and manly character.

Medal of Honor

OF THE BOARD OF DIRECTION

Attention is called to an alteration in the terms of award of the Medal of Honor of the Board of Direction of The Institute of Radio Engineers.

At its meeting on May 16, 1919, the Board of Direction decided that, in order to broaden the scope of this award and to enable suitable recognition of eminent service in the radio art, regardless of the time of performance of such service:

The award in question may be made regardless of the time of performance or publication of the work on which the award is based.

The other conditions of the award, as set forth in the April, 1919, issue of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, on pages 95 and 96, will remain unchanged.

SHORT WAVE RECEPTION AND TRANSMISSION ON GROUND WIRES (SUBTERRANEAN AND SUBMARINE)*

By

Lieutenant-Commander A. Hoyt Taylor

(United States Naval Reserve Force)

The purpose of this article is to report briefly some of the work done by the writer for the Navy Department along the lines indicated by the above title. No attempt will be made to give a complete history of all work which has been done on ground wire systems, as this would exceed the scope of this report. This paper will deal mainly with the behavior of short waves on underground and underwater systems, but in order to get a logical story of this work, it will be necessary to include frequent references to long wave work.

1. Clark's Experiments

During the month of April, 1909, Mr. George H. Clark, then Radio Sub-Inspector, United States Navy, conducted underwater experiments at the Navy Yard, Washington and at the Navy Yard, Norfolk, Virginia, using two 30-foot (9.2 m.) launches equipped with two insulated wires 17 feet (5.2 m.) long, submerged 4 feet (1.2 m.) below the surface, being connected to the transmitting and receiving apparatus thru a fixed condenser of capacity 0.003 microfarad. These wires were placed at an angle of 180° from each other below the surface, and extended out from the center of the launch. It was possible to receive signals from the Navy Yard Station, Washington, at 12 miles (19 km.) distance. Working with one-eighth kilowatt at 425 meters, between the launches, it was possible to communicate a distance of 75 yards (69 m.). Further experiments were conducted on board a tug boat off Norfolk. Copper plates were attached to insulated wires and one plate suspended over the bow into the water and the other plate suspended over the stern into the water. The wires leading up from the water to the receiving apparatus

* Received by the Editor, January 20, 1919.

were screened by the use of brass pipe and copper mesh. A fixed condenser of the same capacity as the one used at Washington was used in these experiments. Signals from the Navy Yard, Norfolk, were received at a distance of 15 miles (24 km.). All experiments showed a marked directive effect, received signals showing a great decrease in strength when tug boat and launches were lined up within 60° to 90° of the direction of the sending station. It is apparent that these early tests by Mr. Clark, done with crystal detectors, had to be carried out in such close proximity of the sender, that the results were partly due to real wave action and partly due to conduction currents from the sender. Owing to phase differences, these effects conspire in some directions and not in others, thus giving peculiar and abnormal directive effects. Mr. Clark evidently did not use long enough receiving wires for the best results. As the result of his experiments, the Bureau of Steam Engineering concluded that altho underwater work was possible for short distances, it was not promising for long distance work. It was also believed at that time that underwater communication would only be possible in fresh water. The experiments were, therefore, abandoned. The use of submarine sending and receiving wires in the form of a loop will not be discussed in this report, as that should properly be made the subject of a separate report.

2. New Orleans Experiments

On December 1, 1916, Admiral W. S. Smith and Commander S. C. Hooper inspected the system of underground radio reception which had been brought to the attention of the Navy Department by Mr. J. H. Rogers at Hyattsville, Maryland. Mr. Rogers demonstrated that trans-oceanic signals were easily readable on underground wires. At the same time he went out in a small boat on a lake near Hyattsville and transmitted from the boat with underwater wires to a station at his home about two miles (3.2 km.) away. On March 6th, 1917, the Bureau arranged to have Mr. H. H. Lyon, who had been associated with Mr. Rogers, proceed to New Orleans, Louisiana, with the idea of developing this system for Naval use, bearing especially in mind its possible value at distant control stations and for use between submerged submarines and other ships. Mr. Lyon reported to the Commandant of the Naval Station, New Orleans, and started work under Lieutenant-Commander E. H. Loftin, who was then

District Communication Superintendent for the 8th Naval District. The following is a copy of a report made by Lieutenant-Commander Loftin under date of April 14, 1917, to the Bureau of Steam Engineering, on underground radio experiments.

"Experiments have been in progress during the past three or four weeks at this Station to ascertain the adaptability of underground wires for receiving radio signals. These experiments have been conducted by Mr. H. H. Lyon, Radio Expert, assigned to this duty by the Bureau of Steam Engineering, and Chief Electrician C. W. Jordan, United States Navy, under the supervision of the District Communication Superintendent.

"A standard copper conductor of 23,000 circular mils (11.6 sq. mm.) (7 twisted strands) having rubber insulation of about 0.15 inches (0.38 cm.) thickness, has been used thruout. All wires are buried to a depth of about one foot (0.3 m.), the earth surrounding being practically saturated with water at all times.

"One conductor was laid in an approximate northeast and southwest direction, 1,400 feet (427 m.) extreme length, or 700 feet (213 m.) on each side of the receiver. This wire was cut into sections so that the total length, or several fractions of the total length, could be cut in. In addition to this long wire, two short ones, 300 feet (92 m.) over all, or 150 feet (46 m.) each side of the receiver, were laid parallel to and 10 feet (3.1 m.) from it, one on each side. A switch was arranged to connect either one or both of these wires to the receiver.

"For arc reception a 'BA' type receiver and an ultra-audion oscillating detector were used. Two pairs of telephones were connected in series, one pair being bridged with a type 'A' audibility meter. This was found advisable on account of disturbance of wing potential by the meter.

"The following results were obtained with arc signals:—

"(a) 1,400-foot (427 m.) antenna, grounded at the extreme ends to large plates buried in moist earth, 0.0025 microfarad variable condenser in series for tuning.

Date	Station	Wave Length	Signal Audibility	Static* Audibility	Static Audibility, Main Antenna
Mar. 30	Arlington	7,500	1,000	20	......
Mar. 30	Tuckerton	.....	1,000	20	......
Mar. 30	Darien	7,000	250	20	......
Mar. 30	Pt. Loma	7,000	200	20	......
Mar. 31	Arlington	7,500	1,000	50	10,000
Mar. 31	Tuckerton	.....	1,000	50	10,000
Mar. 31	Darien	7,000	1,000	50	10,000
Mar. 31	Pt. Loma	7,000	400	50	10,000

"(b) 1,400-foot (427 m.) antenna, ends not grounded by plates, 0.0025 microfarad variable condenser in series for tuning:

Date	Station	Wave Length	Signal Audibility	Static Audibility
April 2	Arlington	7,500	1,500	0
April 2	Tuckerton	.....	1,500	0
April 2	Darien	7,000	500	0
April 2	Pt. Loma	9,800	500	0

"NOTE—These arc stations could be read without interference from the Station 5 kw. spark set 300 feet (92 m.) away. Interference began when down to about 5,000 meters.

"(c) 300-foot (92 m.) antenna, ends not grounded, 0.0025 microfarad variable condenser in series for tuning:

Date	Station	Wave Length	Signal Audibility	Static Audibility
April 2	Arlington	7,500	300	0
April 2	Tuckerton	.....	300	0
April 2	Darien	7,000	100	0
April 2	Pt. Loma	9,800	100	0

* Static disturbances here referred to are identical with what are also termed "strays," that is, the sum total of all irregular disturbances of reception irrespective of their (natural) origin.—EDITOR.

"Up to 1,400 feet (427 m.) of wire, the strength of the signal seems to be about in direct proportion to the length of the wire. The directional effect appears to be pronounced. Darien, which is about 80° off the line of antenna, is weaker than Arlington. On the main antenna, Darien is received about 30 per cent. stronger than Arlington.

"The 'B A' receiver, with its static (capacitive) coupler, is not very satisfactory for this form of reception on account of its lack of selectivity and smooth variation of primary inductance.

"At 9:00 P. M., April 7th, it was possible to copy signals from Tuckerton with ease, while static on the main antenna made it impossible to read any arc signals.

"The following results were obtained with spark signals:

"(a) 300-foot (92 m.) wires in parallel, ten feet (3.1 m.) apart, a 0.002 microfarad condenser in series with the primary coil of a Telefunken receiver to obtain 600 meters.

Date	Station	Wave Length	Signal Audibility	Static Audibility	Signal Audibility, Main Antenna	Static Audibility, Main Antenna
April 2	Point Isabel	600	15	. .	. . . .	. . . .
April 2	Tampa	600	200	0	. . . .	3,000
April 2	Port Arthur	600	150	0	. . . .	3,000
April 2	Pensacola	1,200	20	0	100	150
April 2	Ft. Sam Houston	. . . .	150	. .	. . . .	. . . .

"It was planned to take comparative signal strengths on main antenna and underground antenna during the weekly tests, but discontinuance of these tests prevented this.

"Conclusions. The tests have shown that practically all arc stations in the country can be received with ease tho signals are much weaker than on main antenna, but of particular interest is the fact that when static prevents reception on the main antenna, reception can be continued on the underground antenna. This has even been done during a severe lightning storm, when the main antenna would have been dangerous without grounding. Reception is also directional and permits of avoid-

ing interference to some extent by using a wire off-direction of an interfering station. Construction is under way for putting a conductor in terra cotta piping for comparing results already obtained with a plain buried wire. Plans are also being made for trying out the buried wires for locating a distant control station within the limits of the Naval Reservation. A point can be obtained in the southeastern corner of the Naval Reservation, which is 2,600 feet (793 m.) from the radio power house, and it is now thought possible that at this distance, reception can be carried on while the arc is in operation for stations which are on a line at right angles to the line of bearing of the power house. The long wires are being extended to a total length of 2,200 feet (671 m.). Further report will be made if any decided advantage from the additional length is observed.

"RECOMMENDATIONS. It is recommended that all stations in regions where static interference is encountered be equipped with underground reception wires; long wires, about 1,500 feet (458 m.) for arc reception, and short wires, about 300 feet (92 m.) for spark reception. There should be at least two directions covered in order to be able to receive best from different directions. It is recommended that these underground systems be used in conjunction with elevated antennas. With an elevated antenna, when atmospherics are favorable, greater distances can be obtained, but at the same time, when the elevated antenna cannot get business thru, due to static, such stations within the range of the underground antenna can continue their work."

These remarkably high ratios of signals to strays have not been confirmed by observations by other observers at other places. Concerning Lieutenant-Commander Loftin's conclusions, it may be stated that subsequent experiments at Great Lakes, Belmar, New Jersey, and Tuckerton, New Jersey, showed that the extent to which the ratio of signals to strays is improved by the use of the ground wires, depends upon the conductivity of the ground and that the character of the soil at New Orleans is particularly favorable. Such is also the case at Tuckerton, where the wires were buried in a salt marsh. At Belmar, on the other hand, the land wires, buried as deep even as seven feet (2.14 m.), showed very poor ratios of signals to strays on long waves, altho considerable improvement was noted on short waves. At Great Lakes two installations were made, with two different types of soil, one dry and one wet. Observations were in fair agreement with those obtained at other stations. Concerning Lieutenant-

Commander Loftin's recommendations, it may be stated that the recommendation of 300 feet (92 m.) length of wire for spark reception would be correct only for a certain kind of wire with a certain kind of insulation and at a certain fixed wave length. Later experiments at New Orleans and Great Lakes will make this point clear. The fact that the signals on underground wires are weak can, of course, be compensated for by the use of a regenerative receiver and amplification. Lieutenant-Commander Loftin points out, in another report dated June 21, 1917, that there is an optimum length for underground wires for each wave length and estimates it to be one-fifth of the wave length of the incoming signal. This is, in a way, incorrect, as later experiments showed that the ratio of optimum wire length to wave length depends upon the size of the wire and the nature of the insulation surrounding it. Under date of June 13, 1917, Lieutenant-Commander Loftin recommended that the underground system be used in the new distant control installation at New Orleans. Under date of August 14, 1917, the Bureau informed the Commandant that preliminary experiments by the Naval Radio Laboratory at the Bureau of Standards seemed to indicate that strays were largely eliminated by the use of heavily insulated wires, rather than by the use of bare wires. Lowering of the insulating resistance seemed to bring in static to a marked degree.

3. Rogers' Experiments

During the latter part of May, 1917, when the writer was District Communication Superintendent for the 9th, 10th, and 11th Naval Districts, he received orders to report for temporary duty at the Bureau of Steam Engineering, Washington, for a conference with Commander S. C. Hooper, in charge of the Radio Division of the Bureau, on the possibilities of subterranean reception of the type which had already been demonstrated to representatives of the Bureau by Mr. J. H. Rogers at Hyattsville, Maryland. Full credit is due Mr. Rogers for having first demonstrated to the Navy Department that effective reception was possible with subterranean wires, both on long and short waves. The failure of Mr. Clark to obtain more satisfactory results in 1909 is unquestionably due to the inferior detecting systems available at that time. Mr. Rogers' work had been done with audion detectors, and he succeeded in interesting the Bureau of Steam Engineering in the practical possibilities of subterranean reception on the theory that the ratio of signals to strays was superior to what would be obtained with

ordinary aerial reception. The extent to which this theory was justified will appear in the progress of this report and subsequent reports in the same series. On the morning of June 1, 1917, the writer proceeded, in company with Mr. Rogers, to his laboratory which was located on a high hill in a large isolated tract belonging to the Rogers' estate. The land is a rocky formation covered with several feet (about a meter) of reddish, sandy soil, which at the time the tests were made, was fairly dry. The surrounding country is heavily timbered. Mr. Rogers had an underground room into which leads were run from various wires which were buried underground at approximately a depth of one foot (0.305 m.) below the surface, and ran out to the North, Northeast, East, Southeast, South, Southwest, West, and Northwest. These wires varied in length from 300 to 1,000 feet (92 to 305 m.). Some were bare wires, some insulated and some were laid in tile for part of the distance. The receiving instruments used in the tests consisted of a loose coupler for long waves, a loose coupler for short waves, suitable loading coils and tickler coils, and variable condensers. The receiving bulb was a tubular audion going by the trade name of "audiotron," and seemed to be fairly sensitive. The apparatus was by no means ideal for the purpose, as certain variations in inductance, which should have been possible, could not be obtained. Tests on short waves in the neighborhood of 600 meters were made to determine whether the apparatus was directive and whether the wires laid in tile, or insulated wires or bare wires were best. Directivity was very evident, the bare wires giving greatest strength of signal. Mr. Rogers had been unable to get any tuning in his primary circuit, but the writer found, that with proper adjustments of series inductance and capacity, sharp tuning could be obtained in the primary, provided that the secondary was loosely coupled. The proper length of the bare wires for short waves was not exactly determined, but indications were that it lay between 300 and 500 feet (92 and 153 m.) for a 600-meter wave. Similar observations were made on long waves, signals from New Brunswick being received with very satisfactory audibility, the wires showing a marked directivity, altho not so much as in the case of short waves. New Brunswick's wave at that time was 8,600 meters. It was evident that a wire 1,000 feet (153 m.) long was not long enough to get the best possible signal on New Brunswick. It was not possible to make accurate observations on the ratio of signals to strays, but it appeared that the strays were eliminated more completely

on short waves than on long waves. It was not possible to copy trans-Atlantic signals on account of the strays being too violent on long waves. The wires were usually used in pairs in a straight line, one wire being connected to the antenna post of the receiver, and the other to the ground connection of the receiver. Further tests were made using a ground connection against various wires. The ground connection consisted of a pipe driven deep into the ground. Results were not very satisfactory, the directivity being less marked and the signals much weaker. In view of subsequent experiments it is evident that the ground connection was not a good one. Mr. Rogers stated that he had transmitted on short wave lengths with 0.5 kilowatt power input from a small station at his house, but had, so far, been unable to receive the signal more than a few miles away.

4. Great Lakes Experiments

As the result of this test with Mr. Rogers' apparatus, the writer recommended to the Bureau of Steam Engineering that the subterranean method of reception be given a thoro scientific investigation and that the Bureau equip a small portable laboratory at the Great Lakes Station for this purpose. This was agreed to by the Bureau, and work was immediately started upon the writer's return to Great Lakes. Considerable delay was experienced at Great Lakes in getting a portable steel building for use as a laboratory, therefore, on July 10, 1917, work was begun on two sets of buried wires, one running in an east-and-west direction and the other in a north-and-south direction, and which were installed directly under the towers for the main antenna at the Great Lakes Station. It seemed worth while to discover whether underground reception was possible in the immediate vicinity of the antenna and in proximity of a buried counterpoise. One set of east-and-west wires installed directly underneath the towers consisted of 600 feet (183 m.) of number 12 wire,* with the receiving set in the middle. 150 feet (45.8 m.) of both ends of the wire was left bare. The wires were buried three inches (7.6 cm.) deep in dry soil. Another pair of wires, consisting of bare stranded aerial wire, were similarly laid a few feet distant. A single stranded aerial wire, bare, was run in a southerly direction, down a hill into a ravine. The following results were obtained. Position reports were received from vessels east and southeast from Great Lakes, using the east-and-west receiving combination. On several occasions this was

* Diameter of number 12 wire = 0.0808 inch = 0.21 cm

possible when strays were so heavy that it was not possible to use the main antenna. The signals were, however, very weak, much weaker than the corresponding signals received at Hyattsville. Signals from the Naval Station at Ludington, Michigan, 130 miles (209 km.) northeast of Great Lakes, could not be received on any combination of wires. Signals from the Naval Station at Milwaukee, 50 miles (81 km.) to the north, were received distinctly on the east-and-west wires, but were many times weaker than on the regular antenna. Milwaukee's signals were, however, very satisfactory on the south wires used against either the east or west wire. These experiments were all on 600 meters. Arlington's 2,500 meter spark was copied without difficulty on east-and-west wires. Two additional wires, 500 feet (153 m.) long, were laid in an east-and-west direction, and various long wave signals were copied. It was possible to leave these wires connected and the receiving set in operation while sending was going on overhead on the aerial at 600 meters spark. It was not possible to continue reception on account of the loud interference, but the reaction between sender and receiver was not sufficient to cause any damage. Great Lakes had two antennas, one under the other, one with a free wave length of 500 meters and the other with a free wave length of 1,200 meters. It has never been possible to maintain the arc watch on the larger antenna while sending was going on at 600 meters on the smaller one. This indicated possibilities for distant control purposes. It was evident, however, that serious reactions were being experienced fron the overhead antennas and the buried counterpoise. On the whole, better results were obtained on long waves, using the 500-foot (153 m.) wires. Work was, therefore, started at an extemporized laboratory on the sand beach at the foot of the 90-foot (27.5 m.) bluff at Great Lakes. A tent was erected on the beach and long and short wave receivers with amplifiers were installed; and work at the radio station proper was abandoned, except on the 500-foot (153 m.) east-and-west wires which were used for the reception of long waves as soon as it was found that both Arlington and San Diego could be copied on the ground wires with greater accuracy than on the main antenna. Under date of August 14th, it was reported to the Bureau that several receiving sets could be connected simultaneously to the same pair of ground wires, without interference, and without the tuning of the primaries being in any way interfered with. In continuous wave reception, beat tones, of course, result, but unless the waves are too close together this can readily be avoided. It

must be noted at this point that the writer's experiments at Hyattsville showed that it was only possible to tune the primary when a series condenser was used. Wave meter tests made at Great Lakes showed that as far as the primary tuning was concerned, it was determined wholly by the primary inductance and the series condenser and that with wires over 100 feet (30.5 m.) in length, the length of the buried wires had practically no influence on the tuning of the primary. In other words, any ground systems beyond 100 feet (30.5 m.) in length would tune for a given wave at the same primary setting. This would seem to indicate that the ground wires themselves formed an aperiodic system. This is also checked by the fact that multiple reception is possible on long waves.

5. Experiments on the Beach at Great Lakes

The wires on the beach were 90 feet (27.5 m.) below the base of the radio towers and 900 feet (275 m.) distant from the nearest tower. The receiving set in the tent was 20 feet (6.1 m.) from the water's edge. The wires laid out first were all 300 feet (93 m.) long each way and ran approximately north-and-south. The following results were obtained. Very weak signals on short waves were received on bare wires laid in wet sand, while somewhat stronger signals were received with bare wires laid in dry sand. Better signals were received with wires which were either insulated or in dry sand for 200 feet (61 m.) and then laid in wet sand the remaining 100 feet (30.5 m.) each way. The best signals were obtained on well insulated wires laid in wet sand. A regular watch was established from 6:00 P. M. until 12:00 P. M. on August 11th, on 600 meters, using a tubular audion and a regenerative receiving set without amplification. The operator was able to copy every ship worked from Great Lakes, altho one of them was 110 miles (177 km.) to the northeast and several others were approximately 50 miles (81 km.) to the southeast. A large number of ship calls were logged, but as the log covers three pages, it is not reproduced. In two instances the operator received messages correctly upon which a repeat was asked by Great Lakes and by Milwaukee. The arc signals received from New Orleans were very strong but those from Darien were weak. This was due to the short length of wire used. It was possible to copy Arlington's 2,500 meter spark on north-and-south wires, altho Arlington's direction from Great Lakes is much more easterly than southerly. It is noteworthy that Arlington's weather reports and "press" were copied with-

out the least difficulty altho Great Lakes, 900 feet (275 m.) away, was operating on the 600 meter spark thruout the press work. It was also possible to copy Milwaukee, 50 miles (81 km.) to the north, on 600 meters, while the Great Lakes arc was in operation on 6,000 meters, radiating 50 amperes. In order to utilize the directivity of the ground wires in such a way as better to eliminate the interference from Great Lakes, the tent with the apparatus was moved 300 feet (93 m.) further to the north, so that a line from the main radio station to the beach station would bisect the wires nearly at right angles. Wires 300 and 600 feet (93 and 183 m.) long were laid in wet sand and a regular spark and arc watch established on the beach. The wires were buried about a foot (0.3 m.) deep. The following table of observations is typical of the results obtained and shows that when Great Lakes (NAJ) was sending on 1,500 meters only 900 feet (275 m.) away, no serious interference resulted on 600 meters. None of the stations observed were of high efficiency, the average radiation not being over 7.5 amperes and the average height to center of capacity about 125 feet (38 m.).

Time	Station	Distance	Audibility
10:50 A.M.	WME	50 miles (81 km.)	5,000
10:54	WLD	130 miles (209 km.)	50
10:56	WFK	175 miles (282 km.)	50
11:25	WDC	Unknown	500
11:37	WDI	35 miles (56 km.)	1,000
11:40	WFX	40 miles (64 km.)	100
11:42	NAJ	900 feet (275 m.)	400
11:43	WHW	120 miles (193 km.)	125
11:45	WLD	130 miles (209 km.)	100
1:45 P.M.	WFE	35 miles (56 km.)	600
1:50	WFH	35 miles (56 km.)	700

As far as handling regular traffic was concerned, the station on the beach was able to do much better work than the regular station, owing to the elimination of strays and to the ability to work thru storms during which the main antennas had to be grounded. Iron pipe ground connections were driven into wet sand at the end of the wires and while various stations were sending the ends of the wires were connected to the pipe. No difference in signal was noted. The strays were slightly worse. The

directivity of the wires on the beach was very marked, signals from the "Essex," distant only two miles (3.2 km.) straight east from the beach station, were received with an audibility of only 2.5. On the other hand the Naval Station at Manistique, Michigan, 265 miles (427 km.) straight to the north, came in at the same time with an audibility of 400 on the same wave length. When the "Essex" moved away to the south, the signals came up gradually in intensity. Simultaneous reception on arc and spark, with the same pair of 300-foot (93 m.) wires was carried on without difficulty or mutual interference.

Tests with Multiple Wires. The addition of two wires on each side separated several feet (about a meter) from the original wires and of the same length, produced no noticeable change in the signals or strays. This point has subsequently been tested many times. The Great Lakes results do not agree on this point with those reported earlier by New Orleans, but it is certain that the use of multiple wires of the same length offers no material advantage.

6. Experiments by the Naval Radio Laboratory

During the summer of 1917, the experiments of Mr. Rogers at Hyattsville were investigated by Doctor L. W. Austin of the Naval Radio Laboratory, Bureau of Standards, and additional experiments were made under Doctor Austin's direction by Lieut. J. L. Allen, Chief Electrician Nicholson, and Electrician Parks at Mr. Rogers' laboratory at the Fish Hatchery near Hyattsville. Wires were run for some distance overland and then into two small lakes. Observations were made on the Eiffel Tower signals, which were continuous wave at 8,000 meters. The signals were weak but perfectly readable with practically no strays, altho at the same time heavy strays were reported at the Laboratory at the Bureau of Standards. At other times, however, the strays were extremely bad. Doctor Austin reported that it seemed likely from the results at the Fish Hatchery as well as from those at the Rogers' station, that severe strays on ground wires might be connected with the drying out of the ground in the hot sun after a rain. Experiments by the Naval Radio Laboratory at Mr. Rogers' Piney Point Laboratory made in the latter part of August and the first part of September dealt entirely with long wave work. They will be reported in a subsequent paper. Of special interest here, as applying also to short waves, is the

fact that they emphasize the importance of adequate insulation and the desirability of having the wires wholly under water or in moist ground, thus checking experiments which were simultaneously being carried on at Great Lakes.

7. Optimum Wire Length

The New Orleans' experiments having indicated that for best results on short waves, the length of the wire should be carefully chosen, experiments were begun on September 1, 1917, at Great Lakes, to determine the optimum wire length for different wave lengths. The first work was done on 600 meters. The wire used was number 12 rubber covered.* Switches were installed at various points along the wire which was 300 feet (93 m.) in length each way, north-and-south, and audibility measurements were taken with various stations sending and a curve was plotted the vertical ordinates of which represented the audibility of the signal at a certain length of wire, compared with the audibility of the same signal on a comparison wire with a fixed length of 100 feet (30.5 m.). The observations were made on Milwaukee, Manitowoc, Manistique (all nearly straight north of Great Lakes), and Frankfort, and Ludington (which lay to the northeast). The curves all show a very sharp maximum at 125-foot (38 m.) length of wire. The optimum length evidently does not depend upon the direction from which the signal comes. Experiments on optimum wire length were continued, and continuous watch was established on the optimum wire length for 600 meters. It was found possible to receive all stations in the Great Lakes district as far as Alpena in the daytime and as far as Calumet at night. Calumet is 335 miles (539 km.) distant. The strays, as a rule, were practically absent. Occasionally loud cracks, widely separated, were received. These isolated strays, altho loud, did not interfere with the reception of signals on account of their brief duration. On two occasions, strays rose to an audibility in excess of 5,000, using two stages of amplification, but even in this case reception of signals, altho a little difficult, was not interrupted, as the strays were not all numerous. On these two occasions it was necessary to ground both the antennas at the main station. When the optimum wire length is used, it is very important indeed that the wires be fully insulated, since grounding of the wire, either intentionally or accidentally, produces a diminution of the signals. If the wires are accidentally grounded at both

* Diameter of number 12 wire = 0.0808 inch = 0.21 cm.

ends, the signals is reduced 50 per cent. of its maximum value. Grounding the wires intentionally or accidentally, decreases the ratio of signals to strays. The decrease is particularly marked when the wires are carefully adjusted to optimum length. For short waves, the length should be within 5 per cent. of the optimum value for the best results. In order to determine whether the optimum wire length was proportionate to the wave length, arrangements were made for the transmission of test signals from the University of Wisconsin radio station, 9XM, using wave lengths of 425 and 1,125 meters. The bearing of 9XM is 30° north of west from Great Lakes, and is 90 miles (145 km.) distant. The optimum length for 425 meters turned out to be 87 feet (26.5 m.) each way, or 174 feet (53 m.) over all which is almost exactly $\frac{1}{8}$th of a wave length. The optimum wire length for 1,125 meters proved to be 202 feet (61.6 m.) each way or 404 feet (123.2 m.) over all, which is $\frac{1}{10}$th of a wave length, but inasmuch as an insufficient number of observations were taken at the peak of the curve, it is likely that further observations would show this value also to be $\frac{1}{8}$th of a wave length. The length found for 600 meters, 125 feet (38 m.) each way or 250 feet (76 m.) over all, is very approximately $\frac{1}{8}$th of a wave length. The shape of the curves determined at Great Lakes indicated a rather abrupt rise of signal strength when the optimum wire length was secured. The curve is, however, flatter for 1,125 meters than for 600 meters; and subsequent experiments on waves in excess of 5,000 meters, and with very long wires, have failed to show any pronounced optimum wire length. Inasmuch as it was suspected that the optimum length depended upon the electrical constants of the surrounding medium, wires were installed at the beach station directly in the water, which in this case was, of course, fresh water. The optimum length turned out to be exactly the same, somewhat to the writer's astonishment, but the signals were nearly 20 times as strong as when the wires were laid in dry sand. Very satisfactory signals were obtained using two wires, one connected to each post of the receiver and both running in the same direction, provided the one wire was in the water or in very wet sand and the other was in dry sand. The waves evidently experience a change of phase as well as a change in their angle of stagger or inclination with the horizontal. Experiments were now undertaken by Ensign A. Crossley, at Great Lakes, under the writer's direction, to determine whether the optimum length depended upon the size of the wires and the thickness of the insulation. It

seemed highly desirable to use high tension insulation for a permanent installation and experiments were undertaken, first with number 14 high tension Packard cable,* for which the optimum wire length was considerably longer, being in the neighborhood of 200 feet (61 m.) for 600 meters. These experiments continued thruout the summer, and Ensign Crossley was able finally to report that the optimum length was inversely proportional to the capacity per unit length of the wire measured against the ground. In other words, with a given size of wire, the thicker the insulation the longer will be the optimum length, and with a given thickness of insulation, the larger the wires, the shorter the optimum length. These observations are very difficult and a matter of much labor to obtain. It is to be regretted that more of them are not available, made by other observers, for checking-up purposes. The optimum wire length for wires laid in reddish clay soil of the bluff turned out to be the same as was previously determined on the beach, nevertheless it was evident that the capacity per unit length of a wire laid in very dry soil is less than when it is laid in very wet soil, therefore, the optimum length should depend on the nature of the surrounding soil. In order to get an extreme case, wires were laid 3 inches (7.62 cm.) deep in very dry sand on the beach and after many failures on account of frequent rains, a series of measurements were taken which indicated pretty definitely that the optimum length for 600 meters for number 12 rubber covered wire† was 162 feet (49.4 m.) each way instead of 125 feet (38.1 m.) each way. It is evident, therefore, that for best results wires should be laid in fairly wet soil or in water. First, because the signals are much louder; second, because the relative suppression of strays is greater, and third, because then the optimum length will remain fixed. If the wires are laid in salt water, care must be taken not to have them too deep unless conditions are such that very high amplification can be used. The signal falls off very rapidly with the depth in salt water, but in fresh water there is, on long waves at least, no measurable falling off in signal strength down to 60 feet (18.3 m.) in depth. The existence of the optimum length is very helpful in making the system much more sharply selective, a feature which is particularly valuable in distant control work. Figure 1 shows two typical curves obtained in a determination of optimum wire length for number 12 simplex cable, the signals in each case being com-

* Diameter of number 14 wire = 0.0641 inch = 0.16 cm.
† Diameter of number 12 wire = 0.0808 inch = 0.21 cm.

pared with those received on 125 feet (38 m.) of rubber covered number 12 wire. It will be noted that 125 feet (38 m.) was the optimum length for number 12 rubber covered wire. The insulation of the simplex cable is approximately twice as thick as that of the number 12 rubber covered, and it will be seen from the curves that the optimum length is also twice that of the rubber covered wire. It will be noted also that when the opti-

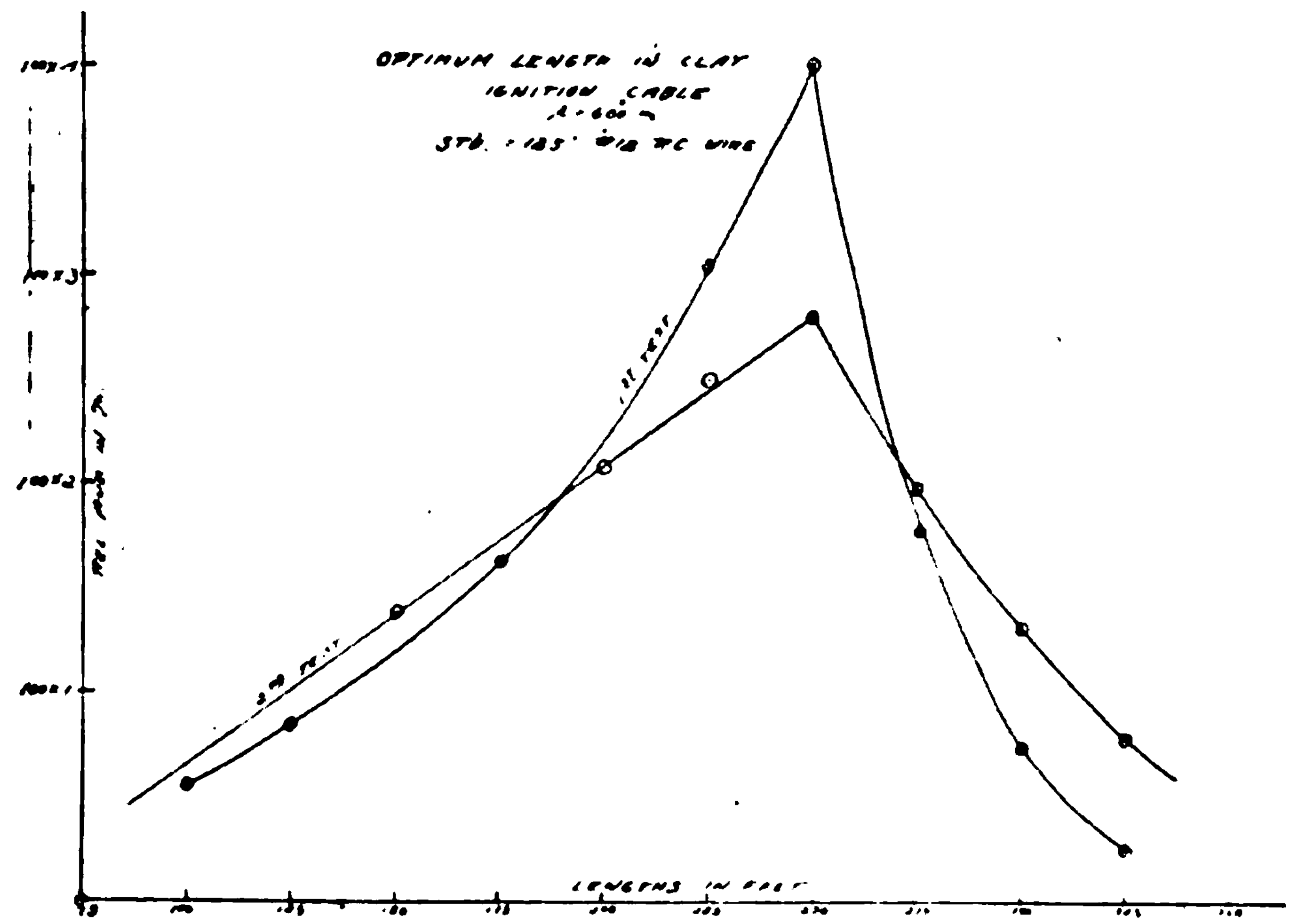

FIGURE 1

mum length is used for each wire, that the intensity of the signal is four times as great on the simplex as it is on the rubber covered. Simplex cable has therefore been used in the final installation, both at Great Lakes and at Norfolk. The optimum length for any given sample of wire may be predetermined approximately by measuring the capacity per unit length of the sample immersed in water and comparing it with the capacity per unit length of a standard wire the optimum length for which is already known. The optimum length of the sample will be related to the optimum length of the standard wire inversely as their capacities per unit length. This holds good, as has already been pointed out, only when the wire is buried in fairly moist soil or in water.

8. Experiments on the Bluff at Great Lakes

The steel building intended for the radio laboratory having been installed in Camp Paul Jones, one-quarter mile (0.4 km.)

straight north of the radio station, wires were laid in trenches four feet (1.2 m.) deep, it being evidently desirable to get down to the level of permanently moist ground for the sake of getting stronger signals and good ratios of signals to strays. Previously determined optimum lengths of wire were used for 600 meters. The trenches radiated north, east, south, and west. A good ground connection was made by driving an iron pipe down to ground water level. It seems curious that ground water should have been found below four feet (1.2 m.) at the location of Camp Paul Jones as it is near the edge of a 90-foot (27.5 m.) bluff, but such was the case. In ground wire work it is, of course, always possible to use one wire against a good ground connection. It has never been definitely settled whether this gives as good elimination of strays as when the wires are used in pairs in the same straight line. The signal is about 65 per cent. of the strength of the signals which are obtained when the wires are used in pairs. The optimum length is the same. If additional wire is added to the ground wire system, but is not buried or under water, it has very little, if any, effect upon the optimum length. That is to say, the optimum length is the length of the submerged or subterranean portion of the wire. If, as Ensign Crossley's reports indicate, the optimum length is determined by the capacity per unit length of the wire, this is, of course, fairly understandable, as the portion of the wire unburied has relatively small capacity. Since the ground wire system is highly directive, the best standard listening-in arrangement for picking up signals is to use a pair of wires at right angles to each other, with a switch arrangement so that when a signal is found, the proper wire pair can immediately be thrown in. The use of a west wire and south wire, for instance, will pick up signals from stations lying in a general north-and-south direction, or in a general east-and-west direction. Signals coming exactly along the bisector of the angle between the two wires, however, will be eliminated. The best universal listening-in arrangement is to use two wires at right angles in parallel with each other connected to one side of the receiver, the other side of which is grounded.

9. Experiments at New London

During the month of August, 1917, underwater experiments were conducted by Mr. H. H. Lyon at the Submarine Base at New London, Connecticut. The water at this point is brackish. The experiments were mostly on long waves and only a relatively

small amount of data was collected. This showed, however, that the results parallel those obtained at other points. It was noted in a report by Mr. Lyon under date of August 11, 1917, that it was possible to work thru a thunderstorm which appeared to be only three miles (4.8 km.) distant and that it was highly desirable to have the wires fully insulated. It is understood that signals were received up to fifteen feet (4.6 m.) under water, altho this point is not specifically mentioned in this report. One stage of amplification was used thruout the experiments. The antenna wires were 500 feet (153 m.) in length.

10. Transmission on Ground Wires

The failure of the early experiments by Mr. Clark on transmission using submerged wires has already been pointed out as having been due to inadequate detecting apparatus. Mr. Rogers' experiments on transmission were far more promising. The problem was taken up again by Ensign A. Crossley under the writer's direction. The following excerpts from Ensign Crossley's reports under date of January 9th and January 23, 1918, are of interest here.

"It is noted that the use of a series condenser and large inductances are essential. By using the ground and one wire, a wave is emitted whose directive transmitting properties are impaired, while by using two wires the directivity of this system is very pronounced; namely, if we use a ground and south wire, signals are received with a maximum audibility at the radio station which is due south of the laboratory, and if the east-and-west wires are used, we find that minimum audibilities are received at the radio station. Altho comparatively low voltages (1,500 to 5,000) were used on the subterranean wires and no trouble was experienced with their insulating qualities, it is practical to use other systems which necessitates higher voltages, provided slight changes are made to insure sufficient insulation. The foregoing experiments were conducted on wires whose optimum length gave maximum received signals on a 600-meter wave. It may be probable that there is a different optimum length for transmitting on 600 meters, but experiments for this optimum length are impractical at the present time due to frozen ground in this vicinity. Experiments of this nature will be conducted in the Spring."

"Much trouble was experienced in operating the 0.5-kilowatt bulb transmitter, due to frequent polarization of the bulbs. This trouble entailed a delay of five days during

which time different experiments were conducted for maximum radiation. Direct connection to ground wires, using inductance and a capacity in series with one wire, the other side of the circuit being connected directly to ground, gave maximum radiation, but unstable adjustment of bulbs. The main trouble with this hook-up is that when maximum radiation is obtained, the bulbs become unstable and frequently polarized. This polarization necessitates complete re-adjustment of the circuits and consequent loss of from ten to fifteen minutes of time. The inductive coupling as shown in Figure 2. gave approximately 0.5 the radiation, as obtained with the first hook-up. This connection required very accurate tuning of

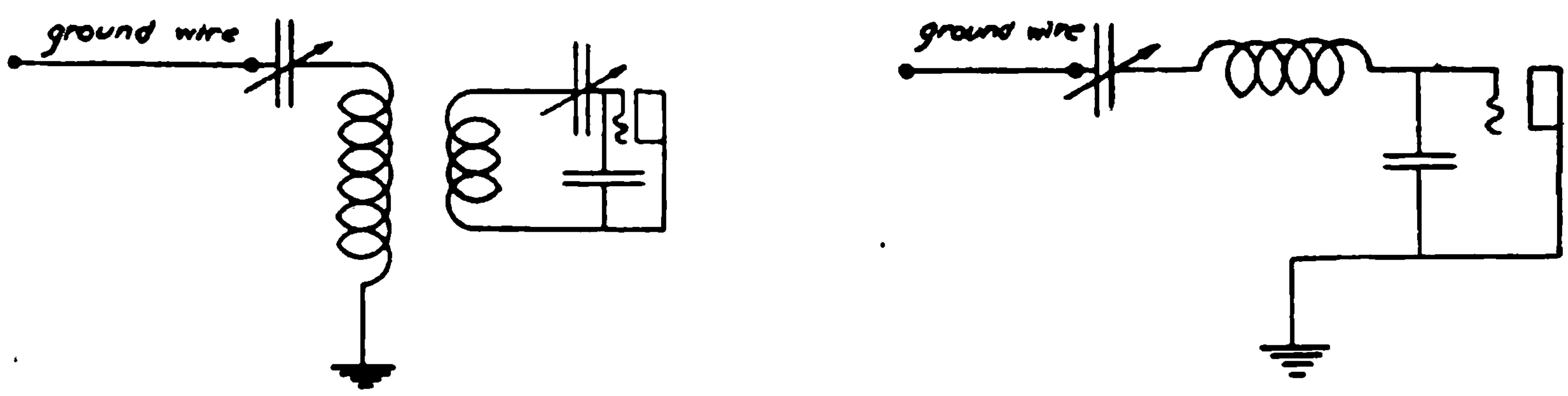

FIGURE 2

condenser before any radiation was obtained. Upon completing these preliminary tests, arrangements were made with Mr. A. L. Howard, of Chicago, who kindly offered the services of his station to conduct tests between the Laboratory and Chicago. One operator was detailed from the Laboratory to assist in overhauling Mr. Howard's receiving station. One afternoon was required to complete this work. That evening the experiments were commenced with Mr. Howard's station, 36 miles (58 km.) from Great Lakes. Wave lengths of 340, 600, and 720 meters were used in this experiment, with respective antenna currents of 0.8, 1.0, and 1.4 amperes. Considerable difficulty was experienced at Mr. Howard's station due to the audion detector being apparently dead from poor vacuum. Two hours were consumed in reviving this detector, after which time the receiving set was in perfect condition. The signals from the Laboratory were heard during the last five minutes of the test. The sending schedule being finished, no further signals were received from the Laboratory. No definite data could be obtained during this test. Arrangements are being made whereby one operator will be detailed permanently at Mr.

Howard's station and daily tests will be conducted between the two stations. If it were possible to use a Cutting and Washington or a Clapp-Eastham 'Hytone' transmitting set, both of which use low voltages, it is probable that better results might be obtained, as the bulb set is very unstable and requires expert manipulation. Efforts are being made to obtain such a set. If this is not possible a Navy portable 500-cycle field set will be used."

Under date of January 31st, Ensign Crossley reported subsequent experiments using the same combination of apparatus, except that the signals from the underground system were compared with those obtained by using the same transmitter on a four-wire antenna 200 feet (61 m.) long, 10 feet (3.5 m.) high at one end and 25 feet (7.6 m) high at the other. A wave length of 720 meters was used on the antenna and the antenna current was 1.8 amperes. On the underground system two wave lengths were used, one of 450 meters and the other of 550 meters. The radiation of the underground system was kept at about 0.5 amperes for each wave. The test was continued for one week, two hours each day. Upon averaging the audibilities of received signals, it was found that the subterranean wires emitted signals that were received at Chicago, distant 36 miles (58 km.), with twice the audibility of the signals from the antenna. The following week the experiments were repeated and showed an average audibility of 1,656 for the subterranean wires and 700 for the aerial. To test the directivity, signals were transmitted on a combination of ground and south wire, and ground and east wire. The results showed that signals transmitted on the ground and south combination were six times as strong as the signals transmitted on the ground and east wire combination. The receiving station in Chicago was due south from the laboratory. So much trouble was experienced with the bulb set that it was replaced with a Clapp-Eastham 0.5-kilowatt "Hytone" transmitting set. While experimenting with this set, it was noted that the addition of extra wires increased the radiation. For instance, if 0.8 of an ampere was obtained by using the ground against a south wire, this was increased to 1 ampere if a north wire was added. When using all the wires, namely, north, northeast, east, southeast, south, southwest, west, and northwest against the ground, a radiation of 1.6 amperes was obtained on 450 meters. The following readings were obtained at Chicago:

Combination	Radiation Amperes	Audibility of Received Signals	Wave Length Meters
Ground with south wire....	0.8	12	450
Ground with north and south wire....................	0.8	20	450
Ground with N, NE, E, SE, S, SW, W, NW..........	1.6	80	450
Regular antenna..........	1.6	60	450

The results indicated in this table are very abnormal. For instance using a north-and-south wire, nothing should have been received at Chicago except that due to a regular antenna effect, due to the fact that the true ground was a few feet (about a meter) below the level of the ground wires. The signals received at Chicago on this combination should have been very small. It later transpired that the south wire had been punctured by the higher voltages of the Clapp-Eastham set. What we had to deal with here was probably a loop transmission effect with the loop underground. The north buried wire constituted the top of the loop and the ground constituted the return or under side. The writer suggested that a counterpoise wire 2.5 feet (0.76 m.) above the ground be tried out against the north wire. This proved to give the best results, possibly due partly to the fact that less strain was placed on the insulation between the buried wire and ground. No ground connection was used with this combination. The signals as received at Chicago were 2.5 times as strong as those sent from an antenna 240 feet (73 m.) long, 9 feet (2.7 m.) wide, 14 feet (4.3 m.) high at one end and 40 feet (12.2 m.) high at the other end, using the same radiation. Ensign Crossley recommended that to obtain more radiation the use of several parallel wires would be of advantage. He was unable to try this out on account of the ground being frozen at the time. It was discovered that the combination using all wires together against the ground had very poor radiating qualities when all wires were properly insulated. During the following week, both the north and south wires again developed a ground, but it was still possible to use them against the counterpoise. The signal was reduced to one-third of its previous value. Experiments were then made with various elevations of the counterpoise above the wires in the ground with

which it was used. It was discovered that the closer the counterpoise was kept to the earth, the greater were the audibilities obtained at the Chicago station, and upon laying the counterpoise directly upon the surface the best results were obtained. Further experiments on transmission were here interrupted by the transfer of Ensign Crossley to Norfolk to undertake the installation of the subterranean distant control station at that point. The writer had already been at Belmar for some time; moreover, the weather at Great Lakes was such as to make it very difficult to continue the experiments.

11. Summary

(a) The history of the Navy's connection with ground wire work, based on the Rogers system, has been briefly outlined.

(b) It has been shown that it is possible to receive very efficiently signals from stations at any wave length, long or short, using submerged or subterranean wires, provided that one or two stages of amplification be used.

(c) The directivity of the wires has been carefully examined and it has been proven that signals coming at right angles to a given wire pair are excluded, while signals coming parallel to the wire pair are received with maximum intensity.

(d) The feasibility of utilizing this system with its highly directive and highly selective properties for distant control work has been demonstrated. A new type of distant control is therefore possible, a type where the control station need be removed only a few hundred feet (50 or 100 m.) from the sending station.

(e) The optimum wire length has been determined to be independent of the angle from which the signal comes and to be, for short waves, roughly proportional to the wave length. The existence of the optimum length greatly increased the selectivity of the receiver. The optimum length has been determined to be independent of the nature of the surrounding medium, if the same is wet or moist. The optimum length depends inversely upon the capacity per unit length of the wire used.

(f) The importance of adequate insulation has been indicated.

(g) The relative advantage of ground wire reception on short waves lies in the ability of the receiving operators to continue to copy messages thru violent storms without danger to themselves and with little or no interruption of traffic. The very great advantage in the suppression of summer strays is noted on all waves but particularly on short ones. The manner of the suppression of strays seems to be not only a reduction in

intensity all around, but a remarkable reduction in the frequency of the strays. Altho the ground wire system appears in many respects to be aperiodic, the fact that an optimum length of wire exists for short waves, stands as a hitherto unexplained contradiction.

(h) Reception in fresh water or wet soil is enormously superior to that of bare wires above or on the surface or in dry soil. As the wire is lowered into water or into wet ground, the signals increase, whereas the strays are reduced. How deep the wires may be buried with advantage is not yet known.

(i) Preliminary experiments on transmission with subterranean wires have shown that transmission with low powers and continuous short waves is possible over considerable distances. The greatest communicating distance obtained was between Great Lakes and Chicago, a distance of 36 miles (58 km.), with 0.8 of an ampere in the underground wires. Interesting possibilities are indicated by these transmission experiments. With specially insulated wires and amplifying receiving sets, much greater distances can undoubtedly be obtained.

12. Theoretical Considerations

Since ground wire reception may be carried out with wires very near the surface of the ground, altho not so successfully as when buried deeper, the phenomenon must depend upon the well known stagger or inclination of the advancing wave front, thus giving a horizontal component to the electric vector parallel to the receiving wires. It is likely that this angle of stagger increased with the penetration of the wave, especially when that penetration reaches a soil of considerable conductivity in comparison with absolutely dry soil. It is possible, of course, that buried wires may also act in a way like loops, the capacity of the wires to the ground furnishing a return thru ground for the lower side of the loop. Reception on ground wires follows the same law of diurnal variation and seasonal variation as reception on an ordinary antenna. It is not believed, therefore, that we have here to do with a separate wave or current in the earth. This point has not been specifically mentioned in this paper, but has been thoroly proven by observations on trans-Atlantic work. The writer is unable to form an opinion as to whether the suppression of strays is due to the material surrounding the wire acting like a Dieckmann cage or to the fact that possibly the origin of the strays is at such a point that their horizontal electrical components are relatively less than those

of the signals. The aperiodicity of the wires is also unquestionably an important factor in eliminating strays. Since the signals received on ground wires are admittedly much weaker than those received on antennas of comparable dimensions, it is evident that transmission on ground wires will be limited in efficiency by the low radiation in horizontal directions. It has, however, the advantage of high directivity, and as shown by Great Lakes experiments, is on low powers, decidedly more efficient than transmission on closed loops. The transmission experiments have hardly progressed to the point where any definite conclusions may be drawn.

SUMMARY: After an historical review of the work of the United States Navy with underground and underwater receiving systems, the author gives data demonstrating the possibility of effective reception on such systems, particularly when using amplifiers.

These systems are found to be directional toward waves travelling parallel to the length of the wire pair. This directional selectivity, which is marked, is applied in control stations for duplex working.

For such underground systems, an optimum wire length for best reception is found to be roughly proportional to the wave length (for short waves) and independent of the direction of approach of the signal. The existence of this optimum length gives further utilisable selectivity. This length is independent of the nature of the surrounding medium and varies inversely as the capacity per unit length of the wire. The wire in question must be well insulated.

Reception thru violent storms, and suppression of summer strays (particularly at short wave lengths) are found.

It is found that lowering such wire systems from above ground into wet soil or into water greatly increases the signal strength and diminishes strays.

Transmission at short wave lengths, over considerable distances, using such systems has been found possible with low power sustained wave transmitters.

DISCUSSION

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which numerous experiments are described
ılity of reception on horizontal antennas on
dry sand, over fresh water and under its surface; the fact is also mentioned that at that time Professor Braun was conducting experiments on subterranean reception in the fortress of Strassburg.

the water, and no systematic attempt seems to have been made to do very long distance work or to investigate the relation between strays and signals.

SIMULTANEOUS SENDING AND RECEIVING*

By

ERNST F. W. ALEXANDERSON†

(CONSULTING ENGINEER, GENERAL ELECTRIC COMPANY)

The first part of this paper dealing with duplex telephony and the "bridge receiver" was printed before America's participation in the war, but the publication of the same was withheld at the request of General Squier and Commander Hooper after a demonstration of the system to these officers in Schenectady. The author also had the opportunity to make a demonstration of the bridge receiver on the battleship "New York" in accordance with the request of Commander Hooper; and it is his understanding that further applications of the system of simultaneous sending and receiving to war ships have been made by the Navy.

The object of this development was briefly to provide means for neutralizing the overwhelming intensity of the transmitted signal so as to make the receiving set sensitive to the faint impulses of the distant signal. Popularly speaking, the corresponding equivalent in sound waves would be to have an ear which could be so adjusted that a person could stand close to a steam whistle without hearing the whistle but listen to a person speaking from a distance of a few hundred feet (about a hundred meters). A successful solution of this problem was found as described. This method of reception which is characterized by a static bridge neutralization may be properly classified as the "bridge receiver."

THE BARRAGE RECEIVER

During the war the same problem presented itself again in a form which called for a new solution. Distances are only relative, and a steam whistle located in Germany might make

* Received by the Editor, February 8, 1919. Paragraphs and figures starred thus: *, were received March 14, 1917. Presented before the Institute, New York, April 2, 1919.

† In connection with the experimental work referred to in this paper, the author wishes to acknowledge the co-operation of Mr. H. H. Beverage and Mr. B. Bradbury.

such a noise that it would completely drown out both in England and France the sound of the voice calling from America. To find a way to counteract such a contingency was seriously considered by the Inter-Allied conference in February, 1918; and appeals for a solution were conveyed to the author by the French representative in this country, Lieutenant Paternot. The solution to this problem which was adopted by the American as well as the French Government after the first demonstration, has become known as the "barrage receiver." This name appears appropriate because the word "barrage" has not only the military meaning which has become so familiar but also the original meaning of toll or stoppage prevention. Thus the barrage receiver may be used not only in time of war to counteract the offensive barrage of an enemy radio station, but it may be used to multiply the number of peaceful communications that may be carried on simultaneously without disturbing each other.

Duplex Radio Telephony

*Everybody who has experimented with radio telephony has undoubtedly observed that the interchange of ideas is not satisfactory if it is necessary to manipulate a switch of some kind in order to change the equipment from sending to receiving. Even if an automatic device is used for performing the change-over, the two parties are apt to say "hello" simultaneously, then wait for an answer simultaneously, then say "hello" again, and finally give it up in despair. It can, therefore, be said that one of the most important problems from the point of view of making radio telephony practical and useful for the general public is to devise a simple method of duplex operation, whereby the speaker is able to hear the voice of the other party in the same way as this is done on the wire lines. In the work that has been done to attain this end several possibilities have presented themselves and have been tried out. It should first be mentioned that Fessenden worked out a system of duplex telephony whereby the same antenna could be used for sending and receiving at the same time. As shown by the patent records, this was accomplished by a system of neutralization in the receiving circuit whereby a high degree of selectivity is attained between the sending and receiving wave lengths. In deciding upon the possible methods of attacking the problem experimentally the above method was left out of consideration on account of the practical difficulties that it appeared to present.

*The first method that yielded practical results was the use of separate sending and receiving antennas, located sufficiently far apart, so that the selectivity of ordinary receiving instruments could be depended upon for differentiation between the wave lengths of the sending and receiving stations. Each pair of sending and receiving stations were interconnected by a wire line and furthermore connected to the exchange of the local telephone system, so that any subscriber on the telephone system could be connected to the radio system. With this arrangement the radio system has the same relation to the subscriber as a toll line. The radio operator takes the place of the toll line operator, and to the subscriber the method of communication is the same as a conversation over the toll line. The diagram of connections is shown in Figure 1. It may be noted that the lines from the sending and receiving stations are introduced in series with the subscriber's line. While a shunt connection can be made which

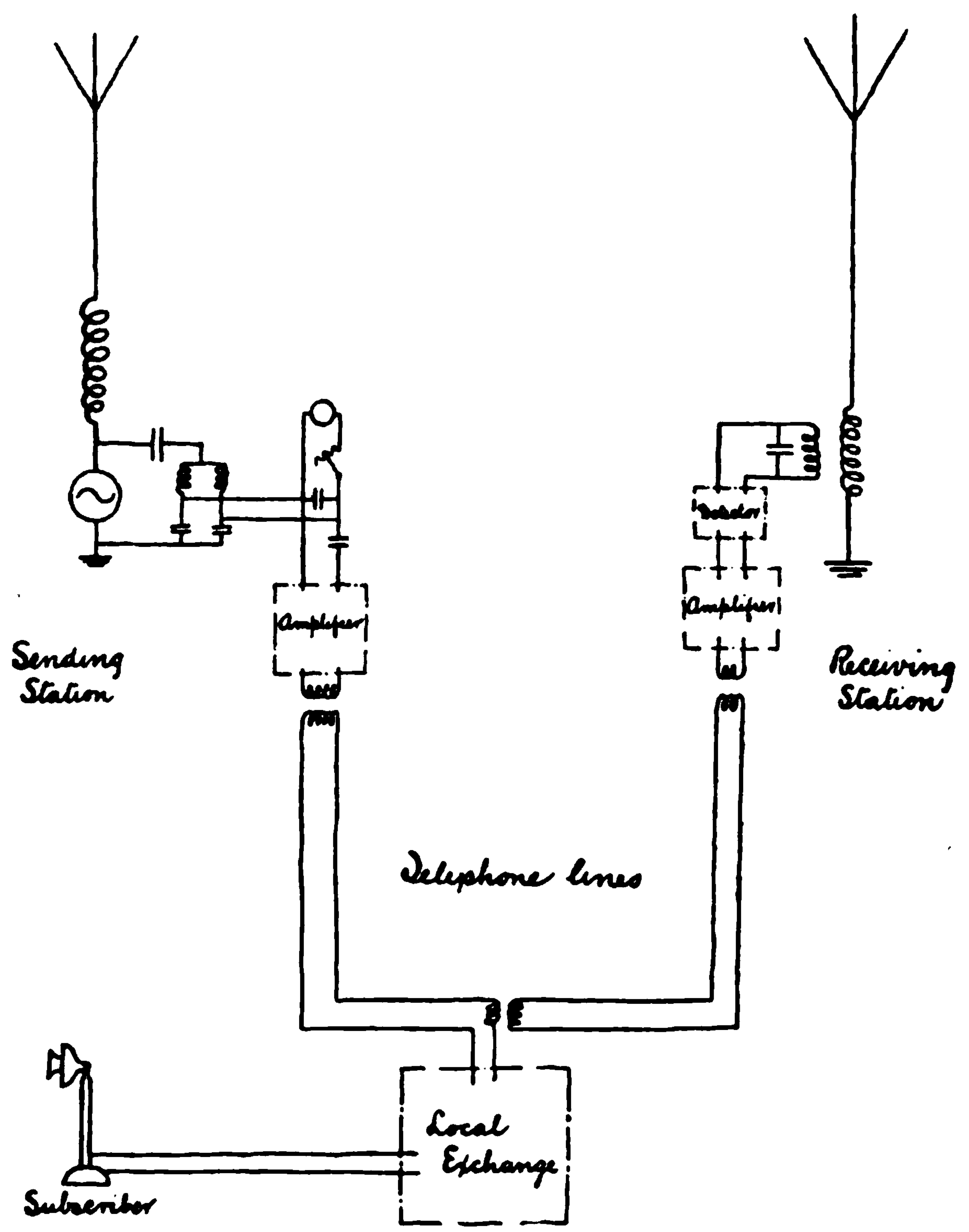

*FIGURE 1—System of Duplex Radiotelephony Connected with Local Telephone Exchange

is theoretically equivalent to a series connection if resistance, inductance, and capacity are carefully equalized, it was found that the series connection could more easily be arranged so as not to interfere with the quality of articulation. The subscriber and the sending station are connected like two ordinary subscribers on a central exchange with the only difference that a transformer with its primary winding connected across the line from the receiving station is, by its secondary, permanently introduced in series with the line to the sending station. A telephone current originating in the receiving station is thus transformed into a current flowing in the closed circuit between the subscriber's instrument and the instrument in the sending station. A telephone current originating in the subscriber's instrument will follow exactly the same path. It thus follows that the current originating in the receiving station will be transmitted by the sending station in the same way as the current carrying the voice of the local subscriber. Consequently both sides of the conversation are transmitted by each sending station and a third party might hear both speakers by tuning in on either of the two wave lengths; this conclusion was confirmed by the tests. Another conclusion can also be drawn from the above reasoning. If the amplification in the receiving station should be made great enough to produce a telephone current in the subscriber's line of greater intensity than the current originally produced by the speaker, this same current will be relayed again thru the sending station and come back to the speaker in intensified form and would again be transmitted from the first sending station. A cumulative effect would thus be created which would result in self-exciting inarticulate oscillations such as may be obtained by holding a receiver in front of a microphone. Any trouble from this source is entirely avoided by keeping the amplification within a certain critical value, whereby the retransmission becomes so rapidly converging as to cause no noticeable interference.

*While the system of duplex radio telephony described will probably prove the most practical for communication over long distances between subscribers of the local telephone exchanges, there are other promising fields for radio telephony for which interconnection with wire telephone exchanges is neither desirable nor practical. Such applications are communication between ships, emergency communication between sub-stations of electric power systems, radiotelephonic train dispatching, supervising stations for forests, and, in general, communication between

isolated settlements in unbroken countries. In all these cases, it is essential that the sending and receiving equipment should be a unit controlled by the same operator. There is, on the other hand, no object in combining the transmitted and received telephone current in the same circuit because the operator may speak into a microphone and receive thru a headphone which two instruments have no electrical connection with each other. The most desirable arrangement under these conditions would, no doubt, be to have a duplex system as indicated by Fessenden, whereby the same antenna could be used simultaneously for sending and receiving. Another possibility of using the same antenna was considered: to use the same set of wires as a loop antenna for the one function and as an open antenna for the other function. For various reasons, however, the following solution, which will be described in greater detail, was found the most practical.

Neutralized Receiving Antenna Mounted on the Same Mast as the Sending Antenna

*If two sets of wires are mounted on the same masts, the radiation from one set to the other is obviously so strong that the overpowering of an ordinary receiving set by the transmitted energy would be almost of the same order of magnitude as if the identical wires were used. The quantitative relations may be better appreciated by mentioning specific figures. In the tests made in Schenectady, the receiving antenna consists of five wires mounted as an umbrella around the main mast, while the sending antenna consists of two wires extending from this mast to another building. The capacity to ground of the sending antenna is 0.003 microfarad, the receiving antenna 0.0011 microfarad, and the mutual capacity such that 10,000 volts on the sending antenna produces 500 volts on the receiving antenna when it is disconnected. While it is obvious that an antenna oscillating with 500 volts continuous waves could not be used with ordinary methods of reception, the system for neutralization which will be described has proven so effective that an ordinary receiving set can be used for receiving signals from such distances as the Pacific coast (2,500 miles or 4,000 km.) without any appreciable interference from continuous wave radiation from the main antenna of 20 amperes and 10,000 volts.

*Two methods of neutralization have been used: inductive neutralization and static (capacitive) neutralization. While both methods have been used successfully, the capacitive

neutralization is much preferable both on account of accuracy of adjustment and simplicity. A diagram of inductive neutralization is shown on Figure 2. The transformer T is used to create a potential of opposite phase to the potential of the sending antenna. The negative potential thus created is transferred to the receiving antenna thru the exposure condenser E. The

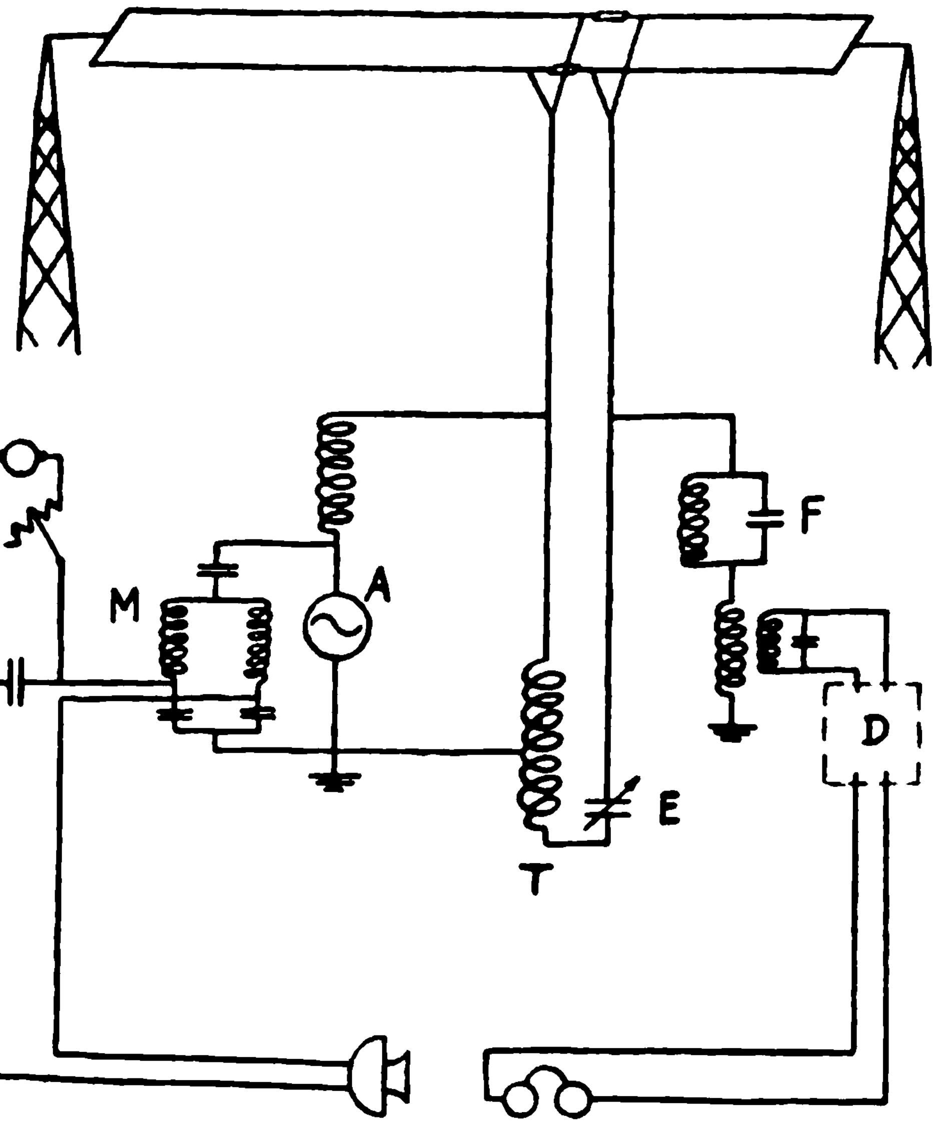

*FIGURE 2—System of Duplex Radiotelephony with Inductive Neutralization

M—Magnetic Amplifier
A—Alternator
T—Neutralization Transformer
E—Exposure Condenser
F—Frequency Trap
D—Detector

negative potential thus impressed upon the receiving antenna thru the transformer and the exposure condenser is adjusted so as to counterbalance exactly the direct exposure from antenna to antenna, thus leaving the receiving antenna at ground potential. The phase relation of the transformer is, however, not exactly 180°, and a residual potential is left on the receiving antenna which is sufficient in most cases to interfere with reception unless further precautions are taken. If, however, a frequency trap F is introduced the neutralization becomes good enough so that an ordinary receiving set can be used.

The arrangement shown on Figure 2 was used to demonstrate duplex radio telephone conversation between Pittsfield and Schenectady (50 miles or 80 km.)

THE BRIDGE RECEIVER

The system of capacitive neutralization is shown diagrammatically on Figure 3. The receiving antenna, A_2, is connected thru a shielded primary loading coil, T_2, to a counterpoise condenser, C_3. This loading coil is coupled aperiodically to the secondary of a receiving set of any ordinary type. The counterpoise condenser is connected thru the exposure condensers C_1 and C_2 to the sending antenna.

The function of the capacitive neutralization can be best explained by showing the diagram as a Wheatstone bridge as in Figure 4. The exposure condensers and the counterpoise condensers form an artificial circuit duplicating the potential drops between the sending antenna, the receiving antenna, and ground. By adjusting the exposure condenser two equipotential points are found between which the receiving set is connected in a manner analogous to the Wheatstone bridge, hence the name "Bridge Receiver."

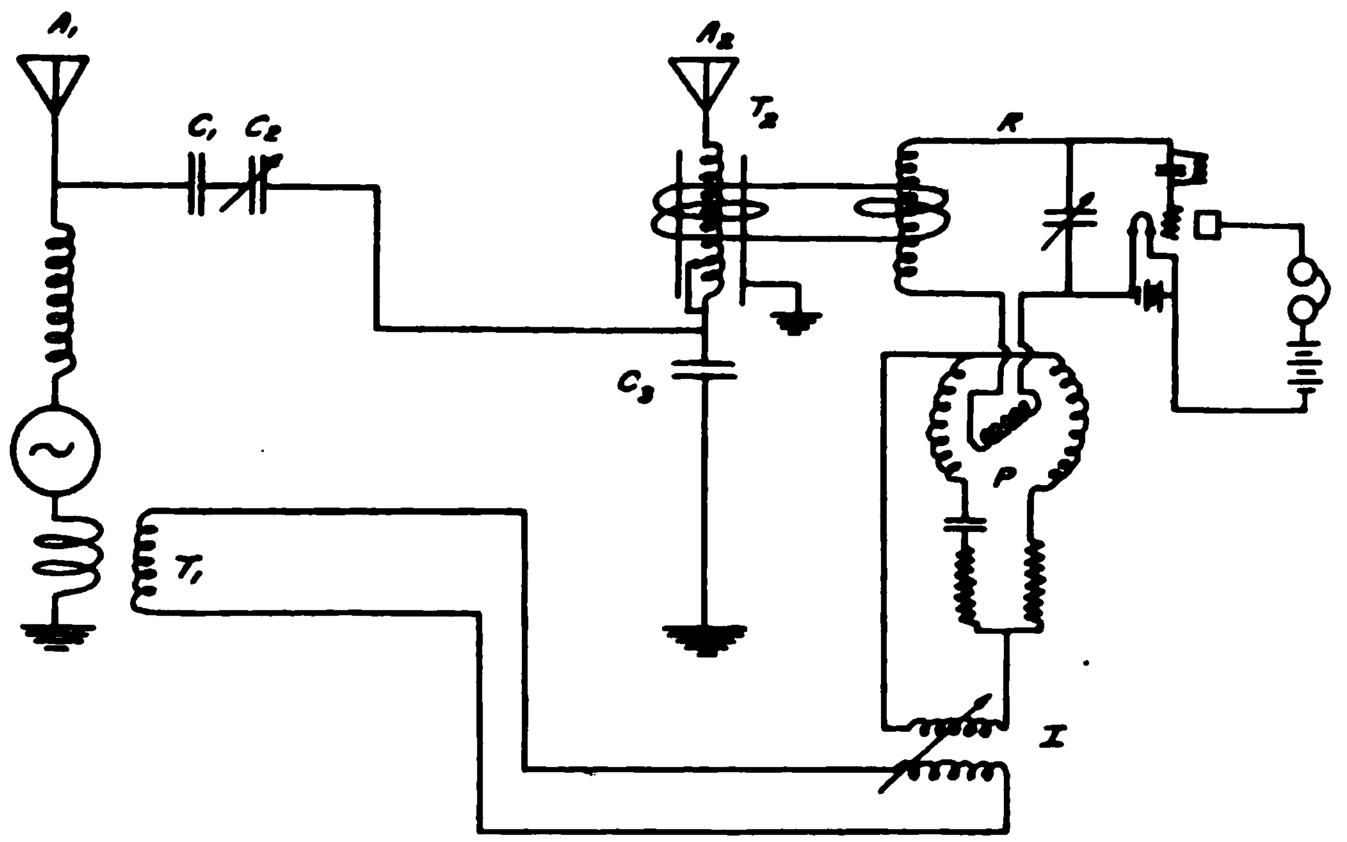

FIGURE 3—System of Duplex Radiotelephony with Bridge Receiver (Capacitive Neutralization)

*The neutralization by this method is so sharp that the influence of the two antennas on each other is reduced much below other sources of disturbance. The principal remaining disturbances caused by the transmitting system are the magnetic strays within the building. In so far as these strays are in phase with the antenna radiation, they are automatically taken care of in neutralizing the antennas. When neutralization is made for minimum disturbance, the neutrali-

zation effect is adjusted so as to compensate the sum of the outdoor and indoor radiation. However, in so far as the indoor strays are out of phase with the capacitive neutralization, a residual effect remains that must be taken care of by other means if it is objectionable. The local magnetic strays cause disturbance principally by interlinking with the secondary loading coil of the receiving set. Evidence of this was found in the fact that the primary neutralization cannot be appreciably

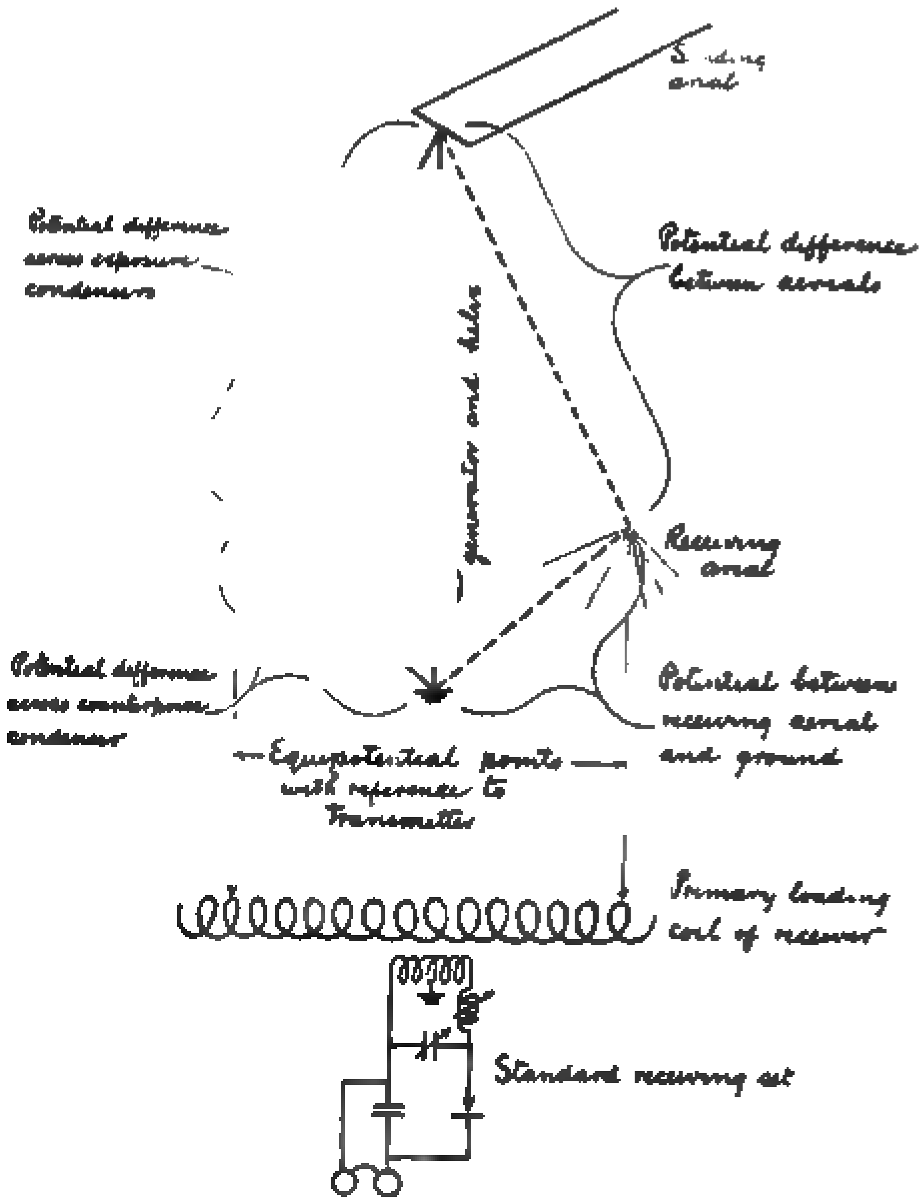

*FIGURE 4—System of Duplex Radiotelephony with Capacitive Neutralization

improved upon by the use of a frequency trap. It has furthermore been shown that the local strays can be effectively neutralized by intercepting the strays on a moderate sized wire loop in the neighborhood of the receiving set and impressing the potential so generated on a little coil mounted with an adjustable coupling close to the secondary loading coil of the receiving set.

Figure 3 shows how the final neutralization is accomplished in a more exact way by a phase rotator, P, coupled to the transmitting set.

Figure 5 shows some simple types of antennas that may be proposed for duplex work. The combination of horizontal and umbrella is the arrangement used for the tests described.

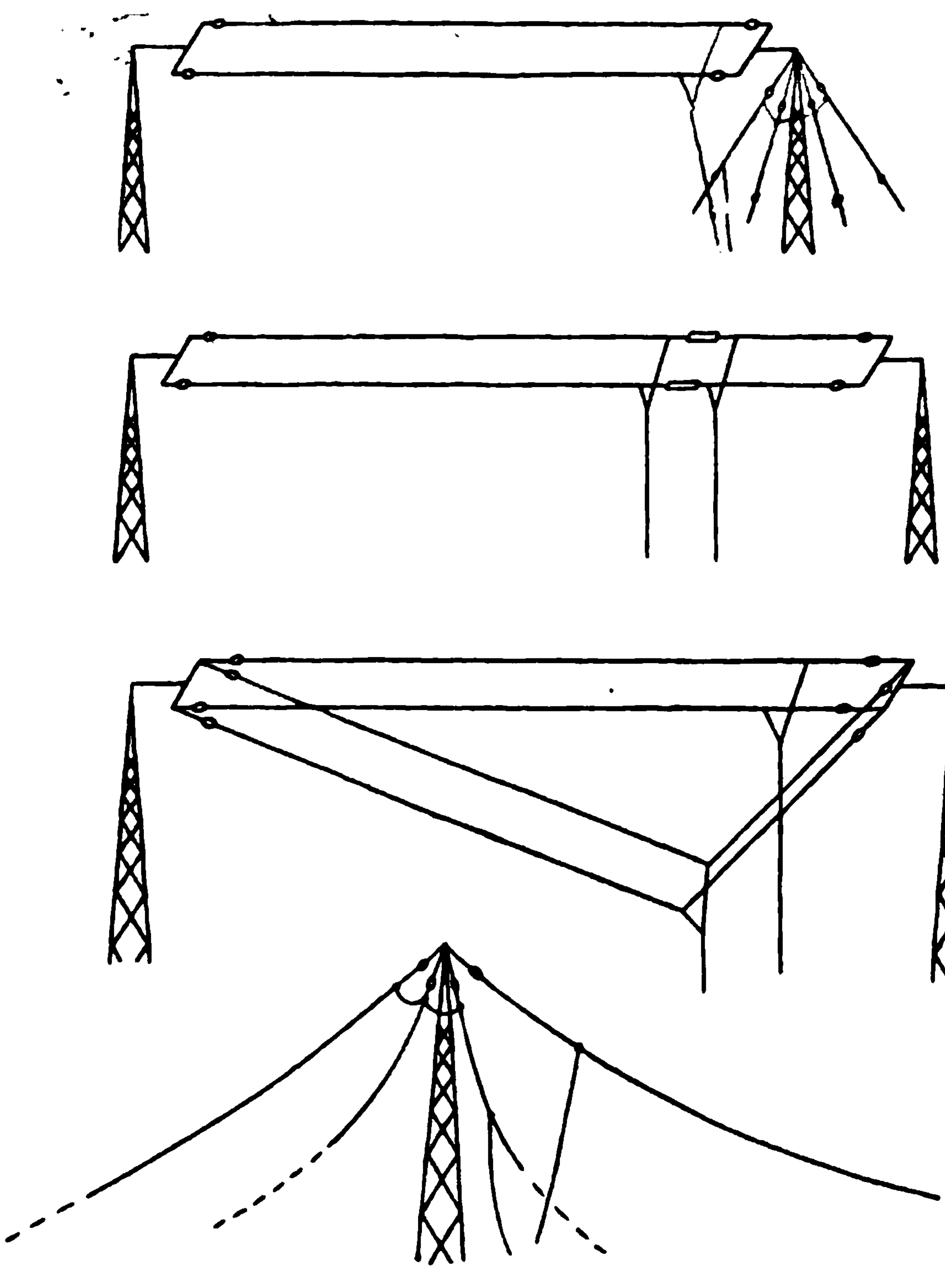

*FIGURE 5—Antenna Systems for Duplex Radiotelephony

Figure 6 is a photograph of a bridge receiving set, consisting of three units. The bridge coupler shown on the left is the shielded primary tuner shown as T_2 in Figure 3. The receiving set shown at the right of the bridge coupler in Figure 6 is an ordinary type of regenerative receiver. The primary of the receiving set is not tuned, but is adjusted to serve as part of an aperiodic coupling between the tuned primary of the bridge coupler and the tuned secondary of the receiving set. On the

right of the receiving set is the pliotron detector unit. The three units shown in the photograph are separate and may be used in different combinations if desired.

FIGURE 6—Bridge Receivin

THE BARRAGE RECEIVER

The barrage receiver is
receiver. The principle of
developed by Bellini and Tosi
Tosi receiver has been used as a direction finder, it has, to the knowledge of the author, not been used to any extent for reception of long distance signals. The Bellini-Tosi receiver is based on the principle of receiving the signal thru two antennas of different characteristics and neutralizes the signals received from one direction by a system of balancing. The principle followed by the author in devising the barrage receiver was—

(1) That the antennas or energy collectors should be aperiodic, because the balance of two tuned circuits is fundamentally very delicate and difficult to adjust for a perfect balance.

(2) That the balancing should consist in neutralizing the electromotive forces in the aperiodic antennas before those electromotive forces have had a chance to create oscillating currents. The phase shifting device should therefore be aperiodic.

(3) The two or more antennas should be of the same character; in other words, it is preferable to balance a magnetic exposure against another magnetic exposure rather than against an electrostatic exposure.

The uni-directional Bellini-Tosi receiver works on the principle that the electromagnetic and electrostatic exposures are 90° out of phase. The barrage receiver takes advantage of

the geographic phase displacement in the wave as it travels over the surface of the earth. In the first barrage receivers which have been installed, the antennas consist of two insulated wires laid on the ground a distance of two miles (3.2 km.) in each direction from the receiving station. It was originally intended by the author to mount the wires on poles, but the easier procedure of laying the wires on the ground was adopted at the suggestion of Lieutenant-Commander A. Hoyt Taylor, and the arrangement has proven entirely satisfactory. The barrage receiving set, photographs of which are shown in Figures 7 and 8, consists of a standard receiving set, combined with a phase rotator set. Figure 8 shows the receiving set proper lifted out

FIGURE 7—Radio Receiving Set with Barrage Section

FIGURE 8—Radio Barrage Receiving Set

The diagram of the phase rotator set is shown on Figure 9. Each antenna is connected to ground thru an intensity coupler, the secondaries of the intensity couplers are connected to the primary of the phase rotators. Each phase rotator is built on the principle of a split phase induction motor or induction regulator. A single phase current introduced in the primary is split into a quarter-phase current which produces the equivalent of a rotating magnetic field inductively related to the secondary. By adjusting the position of the secondary coil, the electromotive force induced in it may be made to assume any desired phase relation to the primary voltage. The receiving set proper when used with the barrage receiver has all the normal characteristics of a standard receiving set. A signal originating in any direction whatever may be neutralized by adjustment of the intensity couplers and phase rotators. This adjustment is very easy to

perform, even by an inexperienced operator, and is perfectly stable after it has been made.

An experimental barrage receiving set was operated for several months of the summer and fall of 1918, about three miles from the New Brunswick, New Jersey, radio station. Records were kept on the reception of European stations during the operation of the New Brunswick station. As the New Bruns-

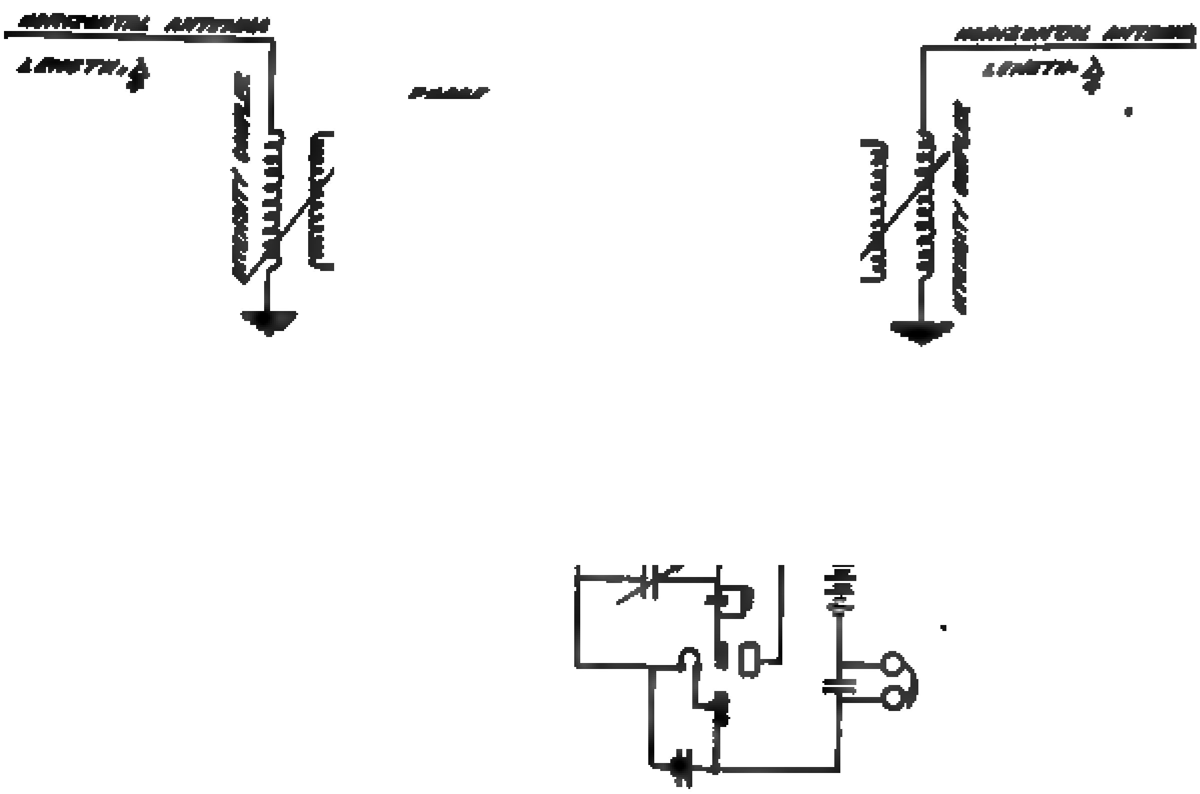

FIGURE 9—Antennas and Receiver of Radio Barrage Set

wick wave is 13,600 meters and the Carnarvon, Wales, wave is 14,200 meters, the reception of Carnarvon was the hardest test to which the set could be put. It was found that in spite of the overwhelming intensity of the New Brunswick signals on an unbalanced receiver, the barrage receiver could be adjusted so that the transmitted wave not only did not interfere with the Carnarvon signals, but the New Brunswick signals could be made entirely inaudible. During these tests it was found that the directive characteristics of the barrage receiver was a material help in reduction of interference by static and strays, as it was found very frequently that solid copy could be obtained by proper directive adjustment, while the signals were practically unreadable with ordinary methods. Statistics of this will be presented in some later paper as the barrage receiver was not originally designed for stray elimination. The improvement of reception of signals by the use of the barrage receiver depends upon the

highly directive qualities of this receiving system. For comparison with other methods of directive reception, a tabulation of directiveness is given. In this comparison the symmetrical elevated antenna which receives equally from all directions is designated at 100 per cent. The percentages of directivity are calculated from the areas of the corresponding horizontal plane intensity diagrams shown in Figure 10. If the directivity represented by the intensity diagrams can be taken as relative measure of the average stray-to-signal ratio, we find that the magnetic loop should have 50 per cent as much strays as the elevated antenna,

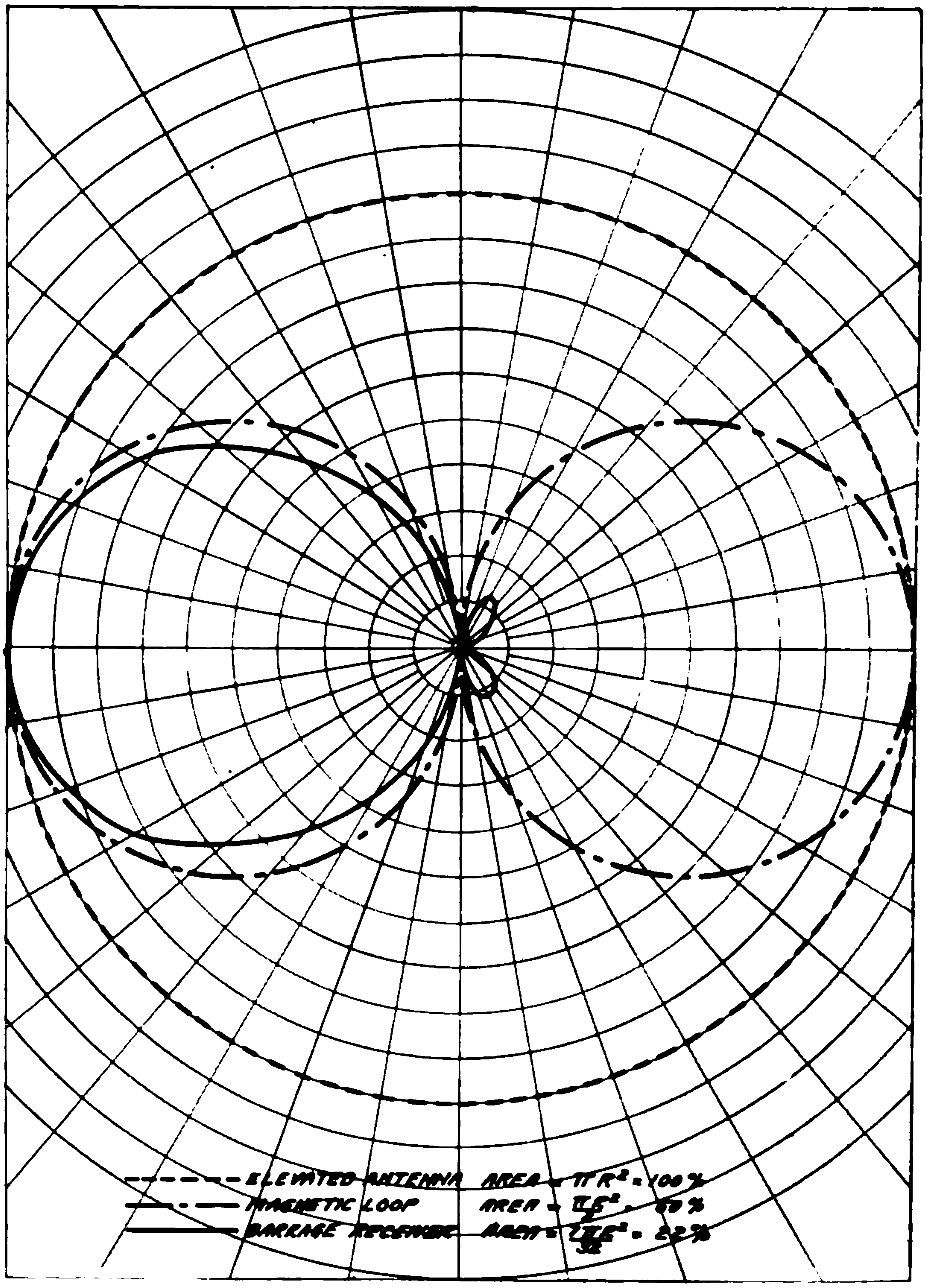

FIGURE 10—Directional Characteristics of Various Receivers

the differential horizontal antenna (Sayville, Long Island), probably about the same as the magnetic loop, and the barrage receiver 22 per cent. Statistics of reception indicate that these figures are reasonably in agreement with facts, when the strays are evenly distributed. When the strays are directive the improvement is much greater.

A rather surprising characteristic was discovered by the use of the barrage receiver. It was expected that this receiver could be used to neutralize signals from all directions except the direction close to the signal to be received. As a matter of fact it was found that interference could be neutralized, originating in the same direction as the signal. This was first discovered in the New Brunswick installation. Signals from San Diego, California, right in line with the transmitting station could be received without great reduction in intensity, while the set was adjusted so as to neutralize the transmitting station. The explanation for this is the fact that in the case of the nearby station, the wave front is curved and the radiation diverging, whereas in the case of the far-away station the radiation is parallel. The receiving antenna covers a space of four miles (6.4 km.), and in this space there is sufficient divergence of the radiation from the nearby station so that an adjustment can be made whereby the diverging and parallel radiation have different effect upon the receiving set. The phenomenon is comparable to the focussing of a field glass on nearby and distant objects. In this case we have a radio field glass of four miles (6.4 km.) in diameter; and, for such dimensions, the focussing effect is sufficient, even at considerable distances, to produce an effective discrimination.

While the barrage receiver was worked out primarily to avoid interference in transoceanic communication, it may be found useful also for the purposes for which the bridge receiver was developed, that is, simultaneous sending and receiving from small shore stations or ship stations. In such cases, it has the advantage over the bridge receiver that it can be used not only to neutralize the transmitting station to which it belongs, but can neutralize interference from any other ship or shore station. By the use of a double set of phase rotators, the barrage receiver may be used to neutralize two stations in different directions simultaneously, and this principle may be carried still further if desired. It is thus hoped that this development will open up new possibilities in dealing with a problem which is perhaps the most important in the immediate future, that is, to meet the

demands on the radio technique for a rapidly increasing number of systems of communication.

January 25, 1919.

SUMMARY: A system of simultaneous reception and transmission for radio telephony is described, together with the reasons for its use. It involves transferring the received speech (from a separate receiving antenna at some distance from the transmitting antenna) to the subscriber's line, and transferring speech originating at the subscriber's station to the radiophone transmitter.

Another type of duplex radio communication is considered, this being based on nearby receiving and transmitting antennas so arranged with their associated apparatus that the receiver and transmitter are in conjugate branches of a Wheatstone bridge. The wiring of the bridge receiver is given and the apparatus shown.

A so-called "barrage receiver" is then described. This is a highly directional combination of aperiodic antennas, with unilateral directional characteristic. When two aperiodic antennas are used, the phase difference of the received currents produced in them depends on the direction of the incoming signals. By phase shifting devices and differential coupling to a common receiver, the signals from any given direction can be balanced out. The wiring and apparatus and its functioning are described.

DISCUSSION

William H. Priess: The ingenious conception of static balance of transmitting and receiving antennas by Mr. Alexanderson, and his disclosure of this invention to the Navy Department in the early period of the War, was responsible for the instigation of a general research at the Washington Navy Yard along the general lines of balancing signal interference. Mr. Israel has covered many of the systems that were tested. I shall attempt to cover some of the general features of the different systems.

At an early stage in the progress of the research it was noted, in some cases, that an absorption of the received signal occurred when the transmitting antenna was connected to the receiving antenna thru the balancing system. An attempt was made to determine the magnitude of this absorption. Reference will be made to the characteristics of five typical systems. (Figure 1 to Figure 5 inclusive.) The method for determining the absorption, due to the balance system, was by measuring the change in audibility of received signal following the change from a simple receiving system to the balance system, in a constant signal sent for this purpose from a distant quenched spark transmitter. The audibility was first checked with the transmitting antenna disconnected. The transmitting antenna was then connected and the best balance obtained. The transmitter was shut down and, with the balance undisturbed, a second audibility reading

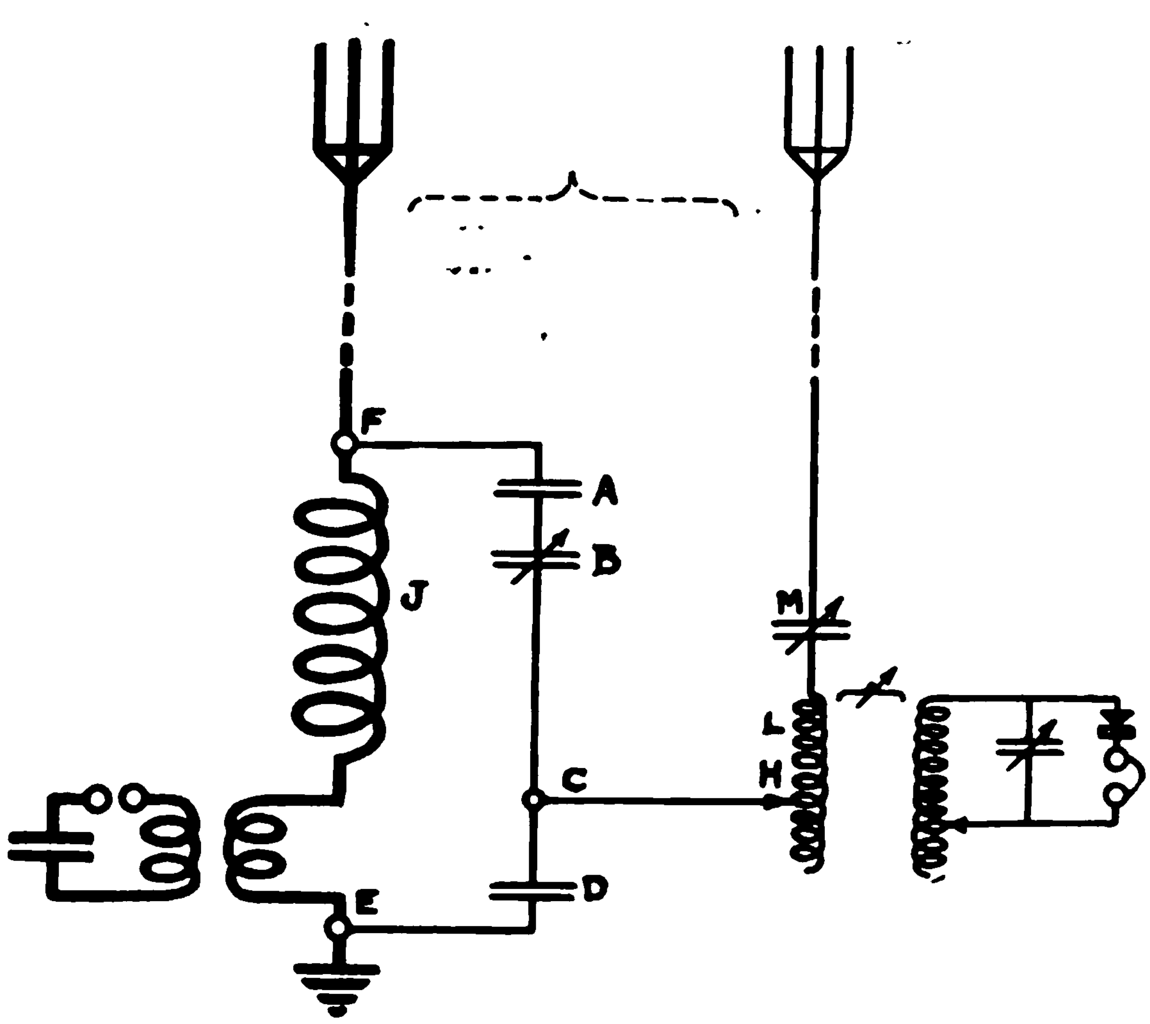

FIGURE 1—Alexanderson Static Balance

was taken. It was found that the systems shown in Figure 1 and Figure 2, as well as a system combining both Figure 1 and Figure 2 (Figure 1, inserting a coil in series with the lead *C H* and coupling it with coil *J*), absorbed about two-thirds of the received signal. Systems 3, 4, and 5, as well as their combinations, showed no appreciable signal absorption, that is to say, showed absorptions less than 10 per cent. It was expected that System 2 and System 3 would suffer equal absorptions. The time al-

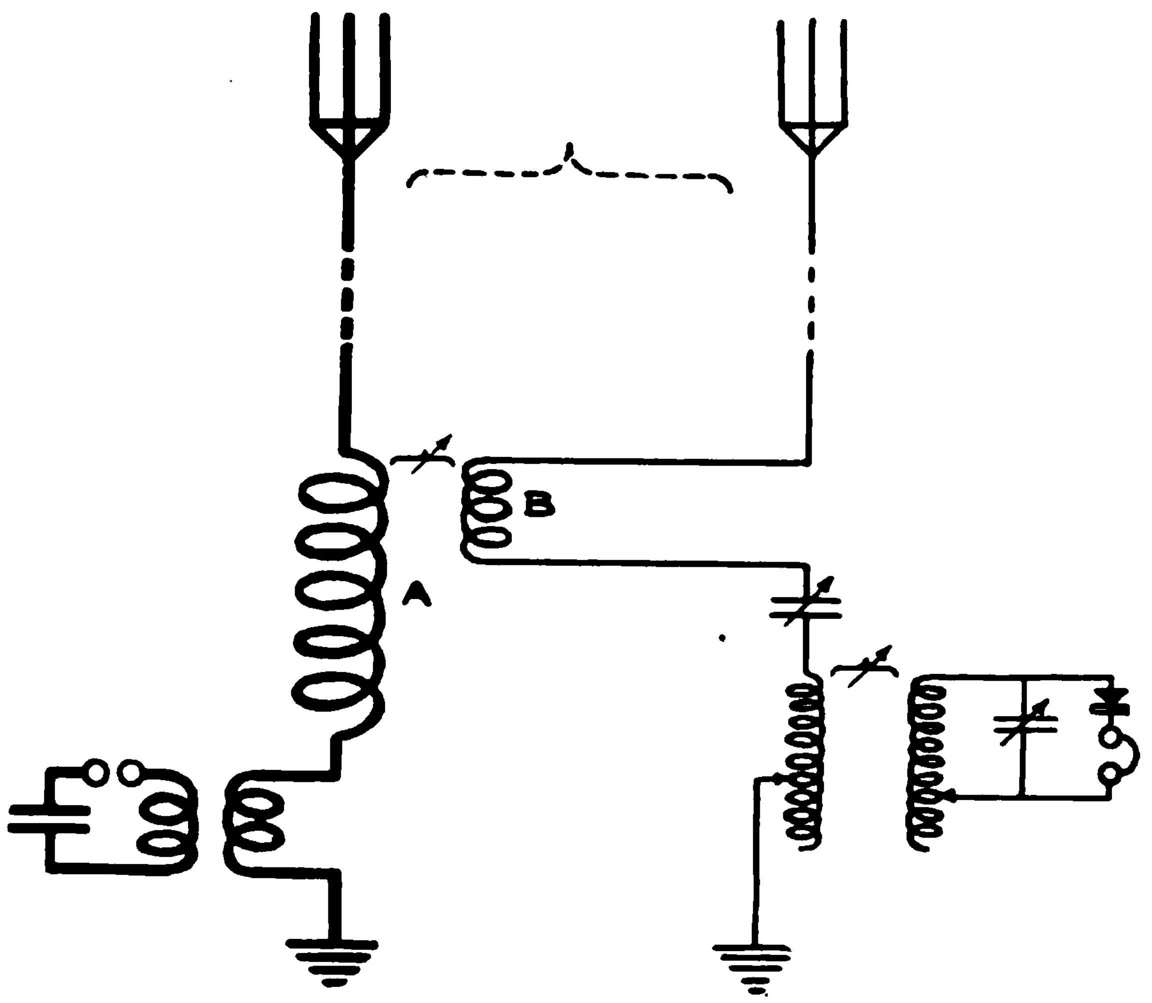

FIGURE 2—New York Navy Yard Magnetic Balance

lotted to the problem was insufficient for thoroly investigating the difference. However, it was found that coil *B*, Figure 3, was opposite in sense to coil *B*, Figure 2, for balance of the two systems. These tests were made with the local transmitter at 1,800 meters and the distant transmitter at 750 meters.

The detuning of the transmitter due to the balance system is an important factor. System 1 caused the greatest detuning and required a process of a series of approximate adjustments between transmitter, receiver, and balancing condensers, that proved very tedious, rendering a wave changer impracticable. Systems 2 and 3 were open to the same objection to a lesser degree. In Systems 4 and 5, this factor did not enter. Combinations of Systems 4 and 5 and uncoupled infinite impedance circuit (inserted for example in lead *AE* of Figure 4), were also free from appreciable reaction on the transmitter.

It was both interesting and annoying to note that minimum residual noise in the telephones and minimum current in the balanced receiving antennas were not in correspondence. With

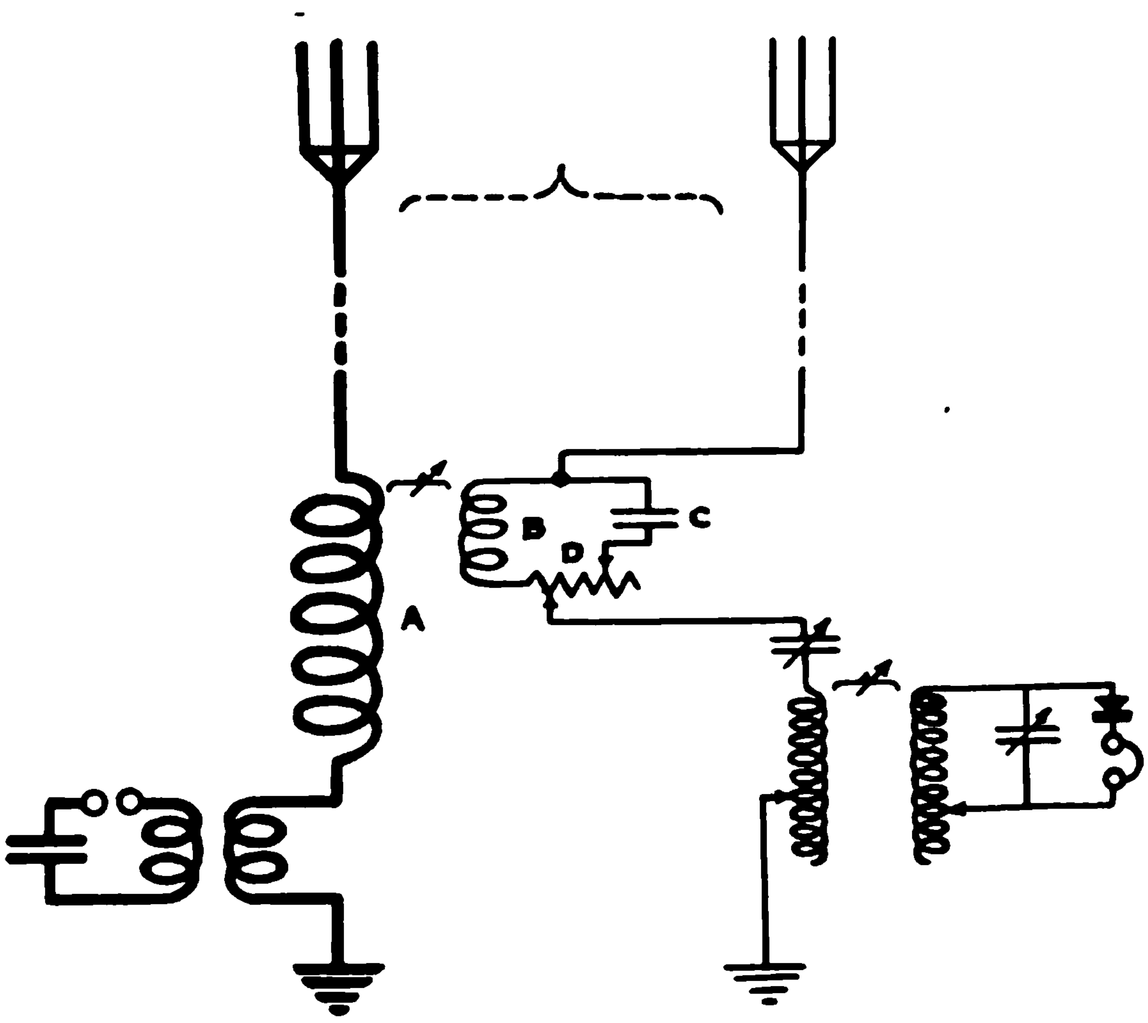

FIGURE 3—Infinite Impedance Circuit

10 amperes in the transmitting antenna at 1,800 meters, and with the receiver at 750 meters; in the first system minimum interference occurred with 160 milli-amperes in the receiving antenna,

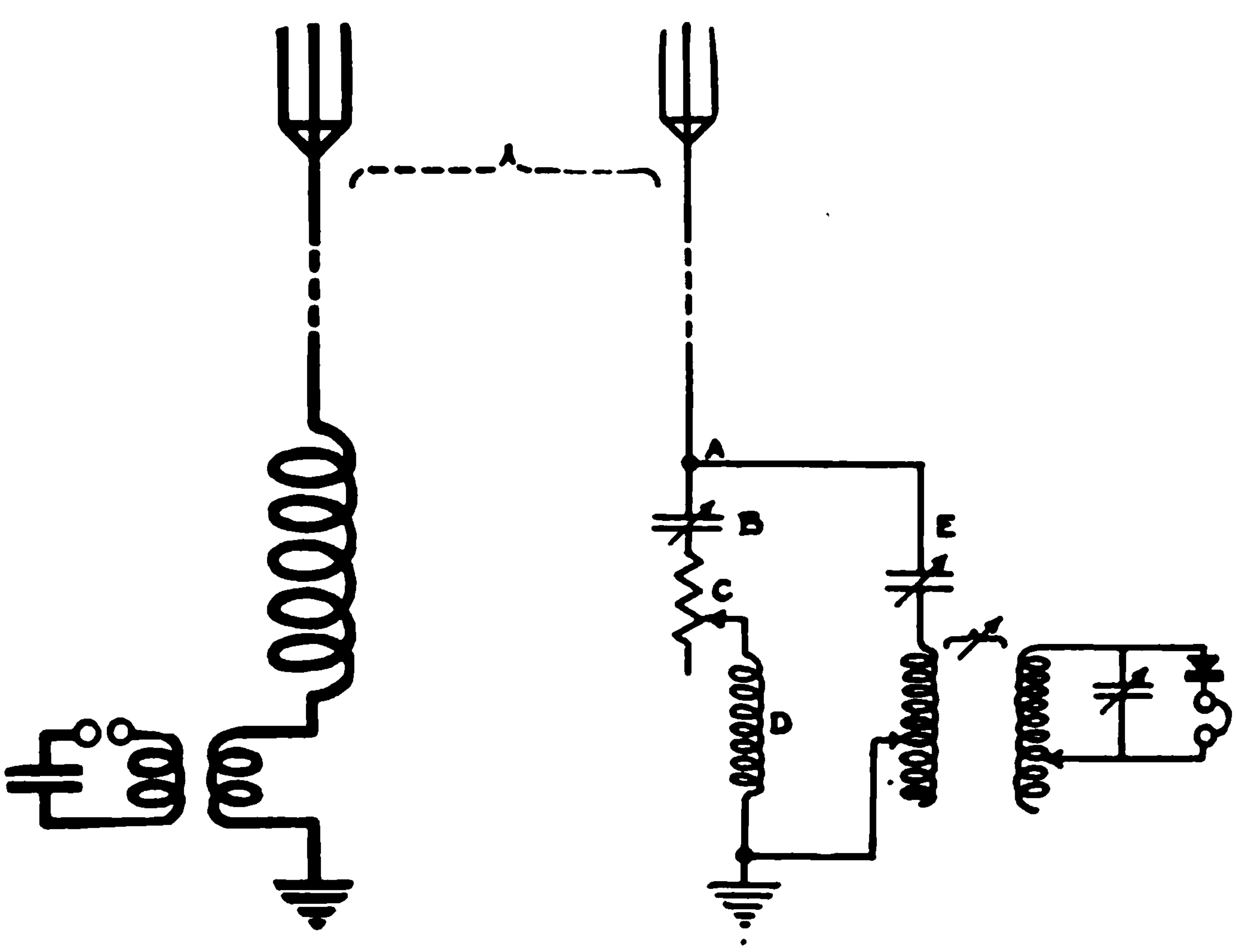

FIGURE 4—Zero Impedance Circuit

altho the receiving antenna could be balanced to 8 milli-amperes. In the second system, minimum interference occurred with 100 milli-amperes in the receiving antenna. In this case, the receiving antenna could be balanced, as the previous one, to 8 milli-amperes. System 5 gave practically a complete balance at the telephones, altho 600 milli-amperes were present in the receiving antenna.

Measurements were made to determine the percentage difference in wave length at which the systems were operable from a practical viewpoint. In the particular case of both the receiver and transmitter in the same room, and both the local and distant transmitters of the quenched spark type, with the local transmitter of the longer wave length; this limit is at about 50 per cent. of the transmitter wave lengths. The detector balance system (Figure 5) proved to be the best combination in this respect. It requires a special receiver primary design to handle the high currents and potentials developed. I made a set of comparative

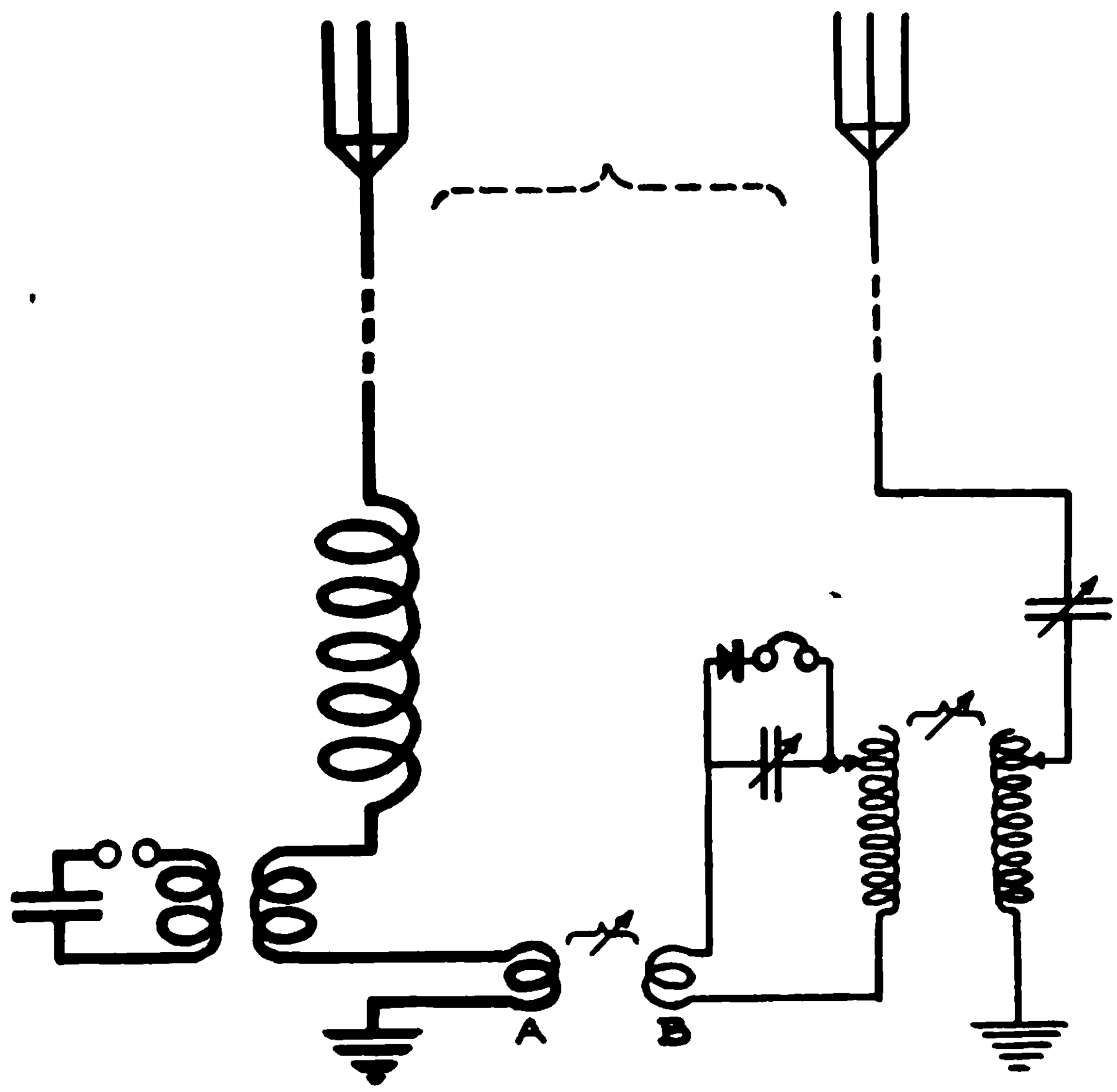

FIGURE 5—Detector Balance Circuit

measurements in an attempt to reach the respective limits of Systems 1 and 5. The antenna systems are shown in Figure 6. The local transmitter was a 5-kilowatt, 500-cycle, quenched

spark set operating at 1,820 meters. The local transmitter wave length was maintained a constant thruout the test. The distant transmitter was likewise a 500-cycle quenched spark set and was variable in both power and wave length. Each set

FIGURE 6

of readings was made with the wave length of the distant transmitter constant, varying the powers of both sets to the limits of the systems. One set of readings made with the distant transmitter at 700 meters, from my notes of June 21, 1917, is the following:

	System 1	System 5
Local transmitter antenna current	7.0 amperes	18.5 amperes
Local transmitter transformer input	0.8 KW.	5 KW.
Distant transmitter antenna current	10 amperes	4 amperes
Primary receiver current at balance	150 milli-amperes	0.78 amperes
Primary receiver potential	Approx. 10,000 volts	Approx. 1,000 volts
Residual noise in telephones	Great	Small
Operators at receiver	M. Kenney, Chief Elec. U. S. N.	Wiseman, Elec. 1st Class, U. S. N.
	J. McDonald, Chief Elec. U. S. N.	Wolf, Elec. 1st Class, U. S. N.
Quality of received signal	Barely readable thru interference.	Very good.

In this test, in System 1, the whole primary of the receiver was at from 10,000 volts to 15,000 volts above the ground, rendering receiver adjustment very difficult. This is apparent from a consideration of Figures 1 and 6. The potential of the receiving antenna due to the transmitter may be represented by some point on coil J, Figure 1. Since the coupling between the antennas is high, this point is shifted towards the antenna end of coil J with a consequent raising of the potential of the whole primary of the receiver ($CHLM$) above the ground. In the

particular antenna combination shown in Figure 6, the capacity of the *AB* condenser combination (Figure 1) was ⅓ the capacity of condenser *D* (Figure 1).

The current and potential design of the receiver circuits are obvious from a consideration of Figures 1-5. However, several general features may be mentioned. In both the infinite impedance and zero impedance combinations it was found that the important condition for minimum residual noise in the telephones was to make the decrements of these circuits approximately equal to the decrement of the transmitter. Extremely high capacity and associated low inductance circuits are to be avoided. Extremely high capacities are necessarily only stepwise variable. Therefore, a continuously variable, extremely small inductance would also have to be used. The brush contact resistance in a continuously variable inductance may introduce a very troublesome factor in raising the resulting uncontrollable decrement of the combination above allowable limits. The circuits should be designed so that a small amount of controlling resistance is permissible. It should be noted that in System 4, variations of resistance *C* between 2 and 100 ohms caused no change in the current in the zero impedance circuit.

The detector balance system requires one variable for control, namely, coupling between transmitting antenna and secondary of the receiver. The coupling between the primary and secondary of the receiver should be pure electromagnetic and fixed at a value that gives maximum audibility with the simple receiver. It is obvious that if this coupling is varied, a variation of the transmitter-detector balance will be required to compensate for the change. An example of satisfactory coupling coils *AB*, Figure 5, used in the research are:

A-4 turns, 3.5 inches (8.89 cm.) diameter

B-9 turns, 3.5 inches (8.89 cm.)

Coils coaxial and separated approximately 3 inches (7.02 cm.)

These coils were adequate for receiving wave lengths up to 1,000 meters when transmitting on 1,820 meters.

The final point I wish to make is the geographical natures of the systems. The receivers of Systems 1, 2, 3, and 5 must be located near the transmitter as they operate on the balance principle. System 4 may be remote as it operates on the filter principle. Zero impedance or infinite impedance circuits, either single or in combination, furnish the most interesting field as their solutions are in the line of general solutions for all cases of simultaneous whether remote or local.

Mr. Alexanderson's second paper on the "Barrage Receiver" is of unusual interest in its vivid depiction of the flexible control of radio frequency phase relations. Looking over some patents of an early date, I found the following combination of tuned magnetic receiving loop with central point grounded. The interesting portion is circuit C for adjusting the phase of the electrostatic component of the wave impressed on the grounded antenna AD, so that both the electrostatic component of the wave impressed on it and the electromagnetic component impressed on the magnetic loop A, are added together in the desired common receiving circuit F. This work was done in 1907 and is the earliest public mention I can find on deliberate phase adjustment in radio frequency circuits. It has apparently not received the general attention of radio engineers for a number of years. This matter of phase adjustment is interesting in view of the papers delivered by Mr. Weagant and Mr. Alexanderson and is merely offered as a suggestion.

John V. L. Hogan: Whose patent was that, do you remember?

William H. Priess; Mr. Pickard's patent of 1908 (876,996).

Another interesting point in connection with the balancing of interference may be made, namely, the balancing of static interference. The ratio of signal to static intensity on a magnetic loop differs from the ratio of signal to static on a straight grounded antenna. The circuit previously shown in Figure 7 provides a system of static balance by amplitude. After adjusting in circuit C the phase of the electrostatic component of the wave received on the grounded antenna DA, so that it corresponds with the phase in circuit B of electromagnetic component of the wave received on the loop A, the two components are added negatively in the receiver circuit F. By varying the coupling between circuits C and F with respect to the coupling between circuits B and F, to the point where the emf. due to static induced across circuit F by circuit C is equal and opposite to the emf. due to static induced across circuit F by circuit B, the total emf, due to static in the detector circuit F becomes zero. However, since the ratio of signal to static is different on the loop from what it is on the grounded antenna, a residual emf. due solely to the signal wave remains in circuit F. The pure signal may be amplified if required without reaction on the system. This is the system of static elimination by

amplitude balance. The apparatus, including antennas, is local and can be used on a ship. Only one adjustment is necessary and that is the relative couplings of receiver and transfer circuits, *F*, *B*, and *C*.

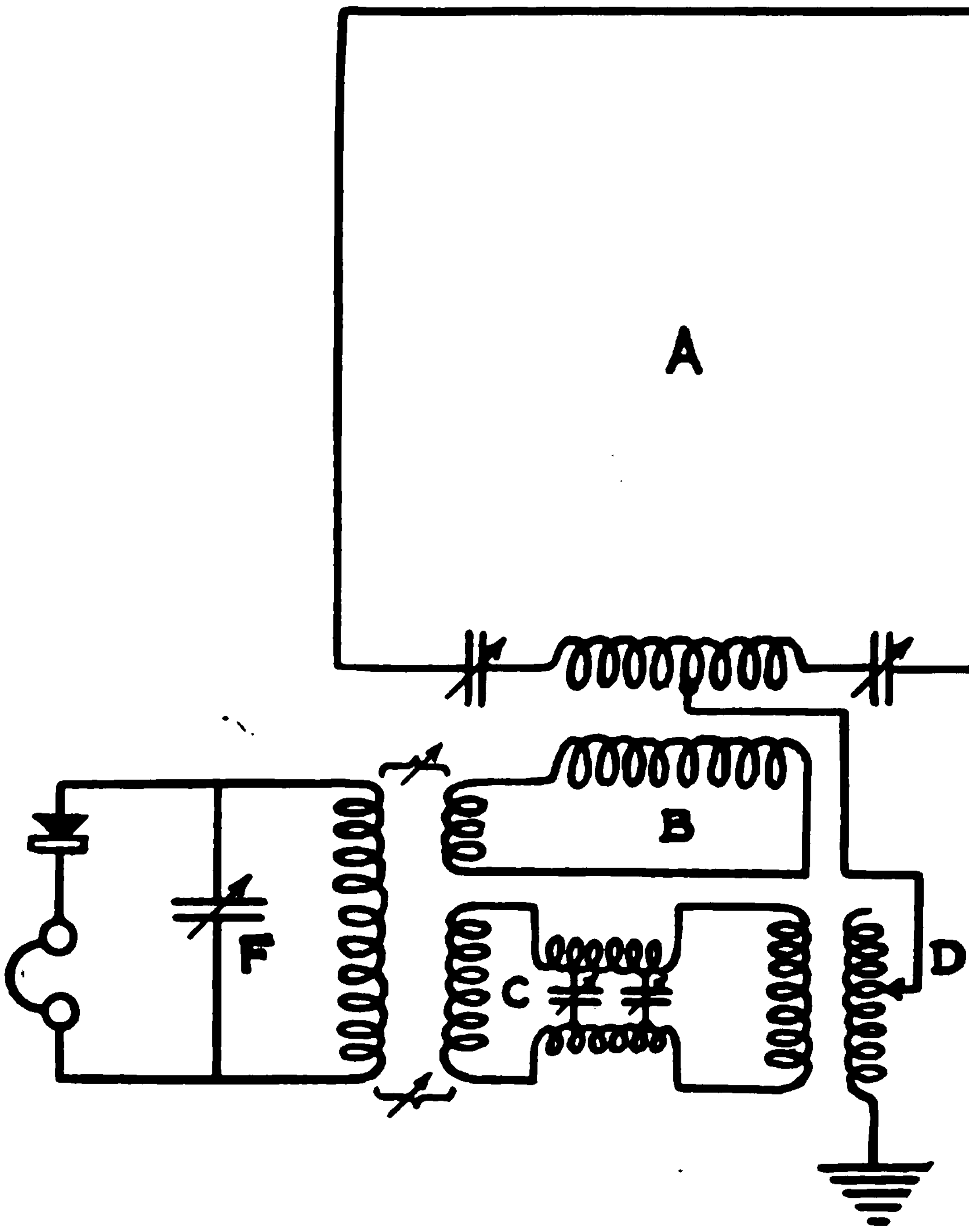

FIGURE 7

Lloyd Espenschied: I should like to ask Mr. Alexanderson if he has available any engineering data on the balance between two closely adjacent antennas which it is possible to obtain experimentally and also that which it is practicable to maintain in service. It should be observed that the two general principles involved in Mr. Alexanderson's work whereby it is possible to exclude the interference from the home sending station are those of selectivity involving a difference in frequency between the sending and receiving transmissions, and of directivity permitting of balance. It would be helpful were it possible for Mr. Alexanderson to evaluate the individual effects of these two methods in separating the undesired from the desired transmission; that is, to what extent and under what conditions may

balance be relied upon and to what extent must balance be supplemented by selectivity Quantitative data of this kind must be had, of course, before duplex radio can be adequately engineered and maintained in service.

Two types of balance are illustrated by Mr. Alexanderson, one in which the balance is effected between the real antenna and a dummy orartificial antenna, and the other in which balance is as between two real antennas. We should expect the maintenance of a balance by means of an artificial antenna to be attended by some difficulty because of changes in the antenna constants caused by varying weather conditions. In this respect, the constancy of balance should be better for the system employing in effect two conjugate real antennas, considering the loop antenna as a form of the latter type.

(After Mr. Alexanderson's answer): It appears then that the balance as measured by current ratio is of the order of a million. It is of interest to note that even allowing for the fact that this result is obtained by refined experimentation, it seems to be much better than the balance which obtains in ordinary duplex operation over land lines. We should, of course, expect to be able to obtain a higher degree of balance in employing a localized structure such as an antenna than is possible under the conditions of wire transmission where the line extends over considerable territory.

Lieutenant M. W. Arps: 1. I notice in the descriptions of experiments made, that Mr. Alexanderson states that the antennas are two miles (3.2 km.) long. Of course, that is impractical on ship work. Antennas 250 feet (80 m.) in each direction are what we have on board ship and what we do work on.

2. Regarding the latter type of barrage receiver, with the experience that the ordinary operator has, it is hard to have him tune two circuits. I should like to know whether the barrage type of receiver is developed to such a point where you could have it on board ship and have the ordinary operator obtain good results.

C. L. Farrand: I am exceedingly interested in the paper Mr. Alexanderson has delivered this evening. The problem of simultaneous transmission and reception and the problem of the barrage receiver are very similar. The latter is better known to me than the first. The general problem involved is producing a receiving system that is responsive to the desired signal and unresponsive to the undesired signal dy differentiating between

peculiar characteristics of each. The desired signal is in all cases that of the distant transmitter. The undesired signal may be from the associated transmitter, which may be very immediate, or at a reasonable distance, an interfering station, static or combination of these three.

Mr. Alexanderson has shown us how successfully a system of this nature can be operated, even when the interfering station was on exactly the same wave length and in the line of direction of the distant station. In the previous discussion, there are mentioned early trials endeavoring simultaneously to transmit and receive, and which have been accompanied by success of greater or less degree. The earliest date of their experiments, I believe, was in 1914. A very early work along these lines was done by Sir Oliver Lodge who, in 1898, took out patents in this country as indicated in the following sketch (see Figure 1), wherein he shows two rectilinear antennas tuned by means of condensers and inductances. Unfortunately, altho he describes the use of the

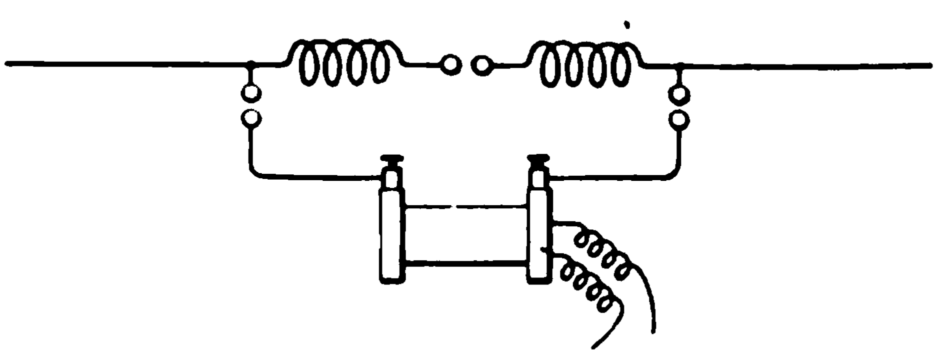

FIGURE 1

system for reception, he does not show the mode of connection, and leaves it to one's imagination that the spark gaps will be replaced by coherers. So it may be considered as stretching the point to consider that his disclosure was suitable for interference prevention. However, an American inventor, very well known to the Institute, Mr. John Stone Stone, in 1901 indicated that he had a very clear conception of the problem by the disclosure in his American patent of a system for simultaneous transmission and reception, somewhat as indicated in the sketch (see Figure 2). Three antennas are shown erected in the direction of propagation of the signal and spaced preferably, the specification says, a total half wave length. The central antenna is used for transmitting, and is spaced a quarter wave length from the adjacent receiving antenna. The receiving antenna is coupled to a common receiver, and in combination is responsive to the distant transmitter and unresponsive to the associated transmitter.

Balancing antennas have been used extensively for different purposes. The Marconi trans-Atlantic stations have used balancing antenna for simultaneous transmission and reception, or "duplexing," as it is termed. The system consisted of a long horizontal antenna, in the direction of reception, usually a single wire several hundred feet (about 100 m.) high and of relatively greater length, for the main receiving antenna. The balancing

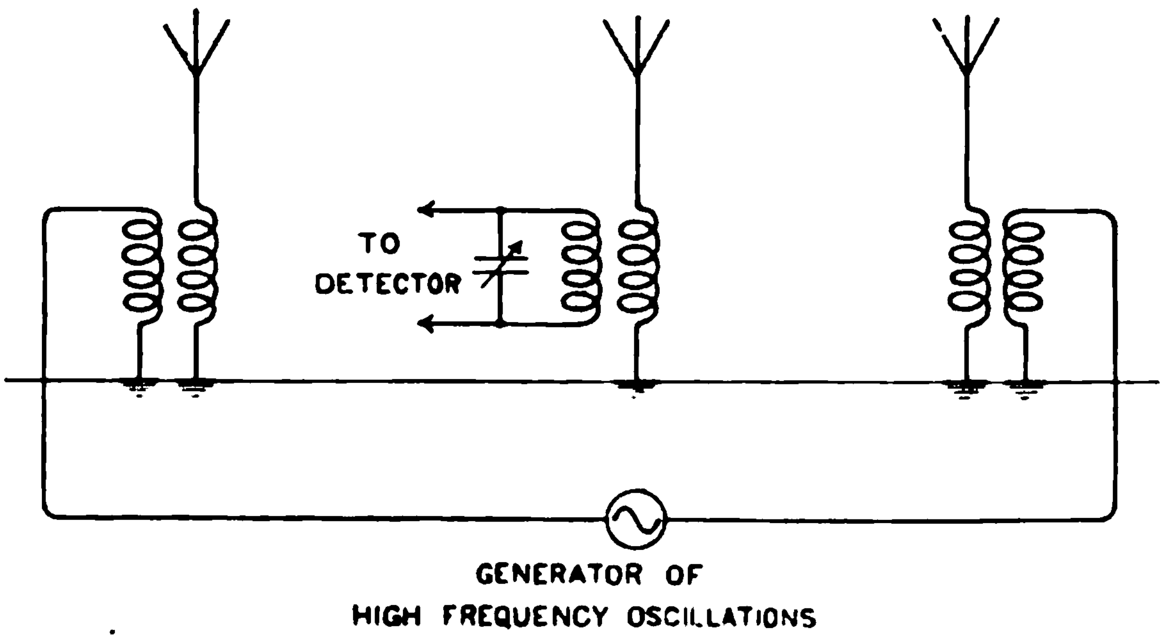

FIGURE 2

antenna was placed at right angles to the receiving antenna, and principally receptive from the associated transmitter, and usually consisted of a single wire less than a hundred feet (30 m.) high and of length equal to the receiving antenna, approximately. The object was to balance out the signal of the associated transmitter. It was considered that the low balancing antenna would receive the signal of that transmitter in greater ratio to the signal of the distant transmitter, than the main receiving antenna. It is rather difficult now to realize that these efforts along the line of duplexing did not accomplish static reduction, but probably it can be accounted for by static reduction not being the object of the experiments, the dissimilarity of the antenna, the relatively poor phase relations existing due to the use of long waves, and the small separation of the electrical centers of the antenna which with the crystal detectors then used did not permit of proper phasing, as sufficient energy could not then be extracted.

Two general methods of attack have been considered in the discussion for reducing interference from signal and static, first is one which Mr. Priess has mentioned, and to which I have devoted considerable attention, dependent upon the inequality of the ratios of the amplitudes of static to signal in two receiving

systems; the second is dependent upon the inequality of phase relation of static to signal in two receiving systems. If two receiving systems could be constructed which would receive static and signal in unequal ratio, a very positive method of eliminating static would result. It seems that the success of the attempts to reduce interference has been due principally to the unequal phase relations, rather than to the inequality of amplitude ratios.

RADIO TELEGRAPHY IN COMPETITION WITH WIRE TELEGRAPHY IN OVERLAND WORK*

By

Robert Boyd Black

(Federal Telegraph Company San Francisco California)

Since the first days of radio telegraphy, numerous attempts have been made by the radio companies to handle commercial business in competition with the land telegraph and cable companies. When there has been any effort at active competition on the part of the wire companies, these attempts have invariably failed. In a few isolated places where wire service was impracticable, radio stations have been in operation for several years and have given a suitable return on the original investment. But it is safe to say that until the introduction of the present efficient type of radio duplexes, competition from the radio companies was not looked upon seriously by the wire companies.

Perhaps if the failure of the radio companies were investigated it would be found to be chiefly a matter of lack of business planning. Most of the half-hearted attempts at competition that have heretofore been made by the radio companies in this field have been carried out with an absolute disregard of expense of operation. Competition to be successful must be founded on a sound economic basis, and with due regard to the operating expenses in comparison with those of other companies in the same field.

Facilities for furnishing reliable and efficient service at a reasonable cost have been lacking in the radio field in the past. Radio stations, for the most part, have been located in inaccessible parts of the cities they were supposed to serve. In some cases, they have been several miles out of the city, which necessitated the relaying of messages over a telegraph or telephone wire, with consequent loss of time, liability to error, and duplication of personnel.

On the radio circuit, the receiver was unable to "break" (i. e., interrupt) the sender, and numerous corrections were

* Received by the Editor, May 1, 1917. This paper was written prior to April 1, 1917.

necessary, caused by unreliable apparatus, atmospheric strays, interference, and other causes. When the sender had finished, it was necessary some times for the receiver practically to skeletonize some messages back to the sending station, in order to secure the scattered missing words from his copy. This resulted in a very ragged looking copy unfit to be delivered directly to a customer, and generally necessitated recopying—a very dangerous practice from the standpoint of accuracy. Moreover, because of the fact that the radio operator has had no means of knowing when he was being called by another station except when wearing the telephones, he has been under the necessity of keeping the head set on constantly or of making "dates" or appointments with the distant stations. Neither of these methods has proven entirely satisfactory from the standpoint of efficiency.

The constant wearing of heavy, tightly fitting head telephone receivers is distinctly detrimental to the health and efficiency of the operator. Besides the danger of ear infection, there is also the liability to other head troubles arising from the constant pressure and strain involved in this method of reception. Furthermore, an operator wearing head receivers is limited by the length of his telephone cords in performing the other necessary duties of a telegraph office. These duties, in a small office, would include waiting on customers, supervising delivery of messages, and attending to the telephone.

Another disadvantage to the radio companies has been the lack of a sufficient number of feeder stations to supply the main trunk circuits. The leading wire companies have offices in practically every town and village, and the greater portion of the traffic which is handled on the trunk lines between the larger cities is relayed from the smaller towns. The radio companies have usually contented themselves with the establishment of stations in some of the larger cities, handling only the business originating at those points. It is manifestly impossible for one company to secure all the business originating in any one city where there are other companies in competition, and it naturally follows that a network of stations should be operated in order to keep the main trunks working to their full capacity.

Lack of efficient organization and thoro training of employees has, in some cases, proven a great detriment to the development of radio in this field. Because of the use of different telegraph codes by the wire companies and the radio companies in the United States, a number of expert Morse telegraphers, who might

otherwise have been available for radio work, have remained with the wire companies. It is hard to train men in a new code in a limited time, and those who have developed into good telegraphers were usually poor electricians, and vice versa. There has been a dearth of fast Continental code circuits in the United States for training operators in high speed work.

All these factors have combined to work against the success of radio in the competitive field; and, coupled with the high cost of operation under these conditions, has invariably resulted in the elimination of the radio companies.

With the recent introduction of greatly improved apparatus for both sending and receiving, it wou'd seem that radio is in a fair way to compete with wire service in practically every department of the work, including the handling of fast press service. The art has lately advanced to a stage of fine tuning, minimum interference, and good results with an amount of power that would not have been thought possible a few years ago.

Multiple radio circuits have been thoroly tried out for a year or more in regular commercial work, and have been found to be quite as efficient and reliable as the wire circuits. It has been demonstrated that messages can be copied directly on the typewriter at a station located in the heart of the city, without undue interference from nearby power and radio stations. The increase in traffic capacity of the circuits and the consequent lower cost of operation has made the future of radio look very promising.

Experiments which are now being conducted, with a view to the elimination of head receivers and the substitution of audible devices for receiving, have shown very good results where the received signals are of reasonable audibility. The great advantage of such a receiving device would, of course, be for calling purposes, but its use in regular work would undoubtedly prove advantageous. With a reliable device of this kind, an operator would be available, when not actually receiving on one circuit, to listen in, or to receive on another circuit, and at all times be able to hear any other station which might call. Following the system of the wire companies in this respect, one operator could cover several circuits, where the amount of business on one circuit would not be sufficient to keep him busy. It would not be necessary to make "dates" with any station. Every station would secure practically uninterrupted service.

From the results actually accomplished to date in radio work and in view of the improvements which are contemplated, a

fairly accurate estimate can be made as to the general organization and conducting of an ideal system of competitive radio telegraph service. Some of the most pressing needs of such a system will be outlined in this paper.

Main Trunks

Between the larger cities there should be established a sufficient number of radio duplexes to take care of all business offered, and to handle it with accuracy and despatch. For the purpose of this paper, these stations will be designated as "relay stations."

The distance by which these stations may be separated is of course dependent upon factors which would be different in different localities. Suffice it to say, that they must be able to carry on fast, reliable communication under all conditions.

"Way" Circuits

From each relay station there should be operated several local circuits, corresponding to the "way" wires of the wire companies. These way stations should be grouped together so that a certain number of them, depending on the business they handled, would send to the relay station on the same wave length. They should each be able to hear every other station in their group in order to prevent interference by simultaneous sending on the same wave.

Relay Station

At the relay stations all business would be handled at the receiving office, which should be located as near as possible to the telegraph center of the city which it served. This would be very necessary in order that the radio company might secure its just share of the available business, and in order to facilitate deliveries from the central location.

The location should be chosen with a view to the erection of an antenna or antennas between two or more comparatively tall office buildings. The best results would probably be obtained by having the space between these buildings and around them comparatively free from obstruction, but this has not been found absolutely necessary in practice. A ground can be made to any convenient water pipe. Very good results in receiving have been obtained from this type of antenna and the expense of erection is slight.

The business office, where messages are accepted for transmission, should have a good ground floor frontage. The operating room should be located in close proximity to the business office, and communication between them should be maintained by a pneumatic tube or similar system.

The operating tables and operators should be grouped together as closely as possible, to facilitate the handling of relay messages. On all duplexes the sender should sit directly beside, or opposite the receiver, and should be provided with a single head telephone or other means of receiving his own "breaks" from the distant station. The sender should also be provided with some device to show that his signals are leaving the sending station properly. There are several ways of arranging a device of this kind, the simplest probably being the introduction of a separate receiving set tuned to the sending wave length. This provides the best possible check on the signals from the transmitting station and the action of the repeating key, and permits instant correction of faults in the transmitter, thereby minimizing the necessity for breaking.

All receiving instruments should be easily adjustable from a sitting position. All variable condensers in the sets should be of small capacity, with long indicators moving in a large arc, so that a maximum movement would be recorded for a moderate change in wave length, and the tuning thereby made finer. The points of maximum audibility should be plainly marked.

All control circuits leading to the transmitting station or stations should be available from each position, preferably by the use of telegraph jack and plug switches.

An easy-running, visible-writing typewriter should be provided for receiving, preferably with all capital letters and without a shift-key. Particular attention should be paid to all time-saving features. Devices for increasing the audibility of received signals should be installed as soon as they are found to be practicable for commercial work.

Transmitting Stations

Transmitting stations should be located a sufficient distance from the receiving office to prevent undue interference. In practice two or three miles (3 to 5 km.) has been found to be sufficient. Normally no receiving should be done at the transmitting station, the whole power output and the regulation of wave lengths being under control of the receiving operator. A

wave changing switch with a sufficient number of points for all necessary changes of wave length should be provided, and these wave lengths should correspond exactly with those of any other transmitting stations in the same city. In case of the failure of one station, another could then be immediately put on the desired wave length and transmission resumed without delay.

Some signal checking device should be provided with the regular equipment at all transmitting stations, which would allow the operator in charge greater freedom for making repairs and attending to other necessary duties about the station. A sparking wavemeter, coupled to the transmitter helix, and consisting of an inductance and a variable condenser tuned to the sending wave length, gives very good results for this purpose. Sparking takes place between the condenser plates, furnishing signals which are audible as long as the set is operating perfectly. The sound may be re-enforced by placing the condenser in a resonator.

In all cases where the power is used intermittently, a considerable saving could be effected by providing distant control of the power source.

Way Stations

Way stations, for the most part, should be established in the smaller towns in the vicinity of the relay stations. Normally, they would transmit their business to the nearest relay station, excepting that which was destined to stations in their own group.

In practice it has been found that altho it is possible for the way station to send its messages directly to its destination, time may be saved by relaying it; for by so routing it work over trunk circuits goes on uninterruptedly.

The duplexing of such stations would hardly pay at present, involving, as it does, the erection of one station for sending and one for receiving, with consequent additional cost for maintenance and operation. These circuits would, in effect, be in multiple at the relay station; and the way station (where one man would act as operator and manager) would be able to "break" at any time to answer the telephone, wait on the counter or attend to other duties. With late types of reliable receiving instruments there should be very little necessity for "breaking" at the relay station. At these the operator would have no outside duties to perform.

In a way station, the complete sending and receiving equipment should be located in a ground floor office, near the center

of the business district, in at least as good a location as that of the competing telegraph companies. In all cases the necessary motor generator set should be started and stopped from the operating table by a remote control. A wave changer should be installed within easy reaching distance, as should the rest of the set. A fast repeating (relay) key should be provided, operated by a small Morse key. A sparking wavemeter should be provided to show that the station was radiating properly.

In the receiving set several pairs of condensers and inductances should be provided, each pair tuned to a certain wave length, and any set instantly available by throwing a single switch. If it were desired to receive on any other than the regular wave length, it would only be necessary to throw one switch, which should be wired up to include the necessary inductance and capacity for the desired wave length. This would save the valuable time lost in tuning for the different wave.

The change from sending to receiving positions should all be done in one switch. The other arrangements should be practically the same as for a relay station.

Operation

In the operation of this system, the way station, before proceeding to call the relay station, would listen in on the regular wave length which had been allotted to its group. If that circuit were busy, the operator would listen on each of the other wave lengths until he found one which was not being used. He would then proceed to call the relay station, which would be provided with a receiving set attuned to that wave length. Upon receiving a reply he would proceed to send his business upon the idle wave length or upon any other which the relay station might designate. With two or three wave lengths allotted to each way station, and these tuned in and covered at the relay station, there should be very little delay on account of the circuits being in use by other stations. These circuits might be designated as "*A*," "*B*," "*C*," etc., to facilitate quick change from one to another.

The relay station normally would call a way station on one particular wave length and then arrange for transmission on another if the regular wave length were busy.

For intercommunication between two way stations, the call would be made on the regular wave length, and transmission effected on one of the auxiliary wave lengths in order to keep the regular wave length clear for calling purposes.

Operating Details

In the handling of this class of business the methods of the American telegraph companies should be followed as closely as possible consistent with proper observance of the rules governing radio communication. The results arrived at by wire companies in the way of improving the working capacity of a circuit are based upon years of experience and probably represent the highest efficiency obtainable in this regard. It has been proven in actual work that the elimination of unnecessary prefixes and other superfluous characters has been a big factor in increasing the capacity of telegraph circuits.

Automatic, high speed work in radio telegraphy at the present time is dependent upon unusually good atmospheric conditions. While some very good results have been obtained in this line at times, the system on the whole is costly and occasions delays which show up badly in comparison with the fast service of the wire companies. A burst of strays lasting only a fraction of a second is sufficient to obliterate a whole word or more of the high speed signals; whereas an experienced operator, copying at the regular rate of speed, would probably be able to fill in the one or two missing letters which a discharge of the same duration of time would cover. Even among the wire companies, the manual system is still recognized as the more reliable for all around work.

The capacity of the circuit, then, depends to a great extent on the skill of the operators employed. There is room for considerable improvement in this portion of radio work in order to equal the work done by the fast wire operators.

Operators

While ability for fast sending in an operator is desirable, it must be combined with other qualities. Good judgment, coolheadedness and ability to read poor copy are just as necessary qualifications in sending as high speed. Generally speaking, a medium-speed, methodical sender, watching his copy closely, eliminating all unnecessary characters, calling attention to misspelled words, avoiding "combinations," and taking especial care in the transmission of code or unusual words, will make better time than the fast, erratic sender. Automatic transmitters, sending purely mechanical signals, have not proven entirely satisfactory for manual reception on account of the sameness of speed which does not allow for the unusual words.

However, the proportion of rapid hand senders among telegraphers is very small, and a hand sender may find his efficiency impaired at any time by operator's paralysis. An instrument called a "sending machine" is being widely used by telegraphers, both wire and radio at the present time. This is an ingenious instrument which works with a side motion, the lever on one side making contact for the dashes, and on the other side starting a mechanical vibrator which makes dots as long as the key is held over. The characters are formed by combining the action of the two. It can readily be seen that there is a great saving in labor when using such a device, and its use practically eliminates paralysis. It is operated with one hand, leaving the other free for marking off and sorting messages. Properly adjusted, a sending machine permits of faster sending and better spacing, in the hands of the average operator, and it will carry thru on any circuit where fast hand sending will carry.

"Combination" sending is a very common fault among telegraphers and one that should be assiduously guarded against. This is improper timing of the dots, dashes, and spacing. This spoils the whole time rhythm, and the receiving operator has the same difficulty in reading the signals as is usually experienced in listening to a stuttering person over a telephone line. Usually the sender is unaware of this fault and imagines he is sending perfect signals. Probably the best method of showing an operator his failings in this respect is to take phonographic records of his sending and compare them with signals which are correctly timed.

Among the commercial operators of the United States, a knowledge of the Phillips Code of abbreviations is coming to be generally recognized as part of an operator's education. This is a system of abbreviations embracing a large proportion of the most frequently used words of the English language, with a complete system of punctuations. It is based on a very comprehensive plan and follows set rules for terminations. The abbreviations are nearly all suggestive of the word which they represent, and are therefore not difficult to memorize.

In practice, the sender simply sends the correct abbreviation for a word, and the receiver spells it out in full and writes it down. Using a typewriter, this is comparatively easy, as the speed of the average typewriter operator is about double that of the fastest sender. This code is used by all American press and brokers' telegraphers, and enables them almost to double their speed. It is also used considerably on the fast commercial wires, and at all times when press is being handled. Knowledge

of it is always a help to an operator, if only for th
gives in conversation in connection with the handlin
It is obvious that if abbreviations are to be used, a regular system should be followed.

The International List of Radio Abbreviations, while forming a handy medium for carrying on a conversation between operators using different languages, is not sufficiently expressive to cover all conditions encountered in telegraph work. Moreover, it is an arbitrary code, not easy to memorize. In the Phillips Code, there are no two abbreviations alike, and if both operators have a good understanding of the code, there is very little liability to error.

A radio duplex, like a wire duplex, is as good as two separate circuits, as long as the number of "breaks" is reduced to a minimum. Therefore both sender and receiver should endeavor to avoid breaks as much as possible. This can only be done by complete co-operation and first class work on the part of each operator. One inefficient man on a duplex can practically nullify the work of the other three.

To the training in this work, and to the general training in concentration, discrimination, and mechanical skill, which only years of experience can give, is due mainly the success in fast radio work of the ex-wire operators who have mastered the Continental code and the necessary technical knowledge. It is admitted among wire telegraphers that it takes on an average of two years or more of all-around experience on the fastest circuits to produce a really first class operator. Even then there are failures, as in every line of endeavor. The equivalent experience on fast radio circuits should give just as good results.

It is obvious that a radio operator must have a good technical knowledge of his apparatus to get the best results. While this knowledge might be subordinated slightly to telegraphic ability in a large office where other expert technical men were available, the knowledge would at all times prove a big asset.

With such an equipment and organization it would seem that a radio company would be in a position to compete successfully with the wire companies in every line of the work. The cost of operation for results achieved should run as low as or lower than that of the wire companies, with their heavy upkeep and right of way costs. The facilities of the radio would, in some ways, be superior to the wire telegraph; and the occasional severe atmospheric difficulties of the radio would be counter-

balanced by the occasional total prostration of the wire telegraph during floods, sleet storms, and the like. In addition, districts could be served where wire service is impracticable because of the physical difficulties which prevent proper maintenance of telegraph lines.

A chain of stations of this type, backed by an efficient organization of employes, would be a valuable asset to the nation and to the districts it served in the event of wire prostration, or national disaster. On this account, it would seem that it would be sound policy for the National Government to encourage and facilitate in every way the extension and operation of such a system.

Probably the nearest approach to the ideal conditions outlined above is furnished by the chain of stations operated by the Federal Telegraph Company on the Pacific Coast. This includes the Los Angeles-San Francisco duplex, and the San Francisco-Portland duplex, and one way "break" systems between Los Angeles and San Diego; Los Angeles and Phoenix, Arizona; and San Francisco and Honolulu.*

The Los Angeles-San Francisco duplex, having been in operation for a period of nearly two years, is probably the best known, and will be described here.

Los Angeles Office

The equipment at the down-town receiving office at Los Angeles consists of two of the latest type sustained wave receiving sets, one normally tuned to receive San Francisco on a wave length of 3,500 meters, and the other normally tuned to receive San Diego and Phoenix on 2,750 meters. The San Diego-Phoenix set can instantly be put in service for receiving damped waves on 600 meters by throwing a single switch.

The antenna for receiving San Francisco is composed of three wires, each 320 feet (100 m.) long, suspended between two office buildings at an average height of 175 feet (53 m.). The San Diego-Phoenix antenna is composed of a single wire, running almost parallel to the San Francisco antenna, and of slightly greater length. No towers are used to suspend these antennas. They are simply swung between the two buildings at the level

* (The distance from Los Angeles to San Francisco is 390 miles, or 625 km.; from San Francisco to Portland, 550 miles, or 880 km.; from Los Angeles to San Diego, 94 miles, or 150 km.; from Los Angeles to Phoenix, 480 miles, or 770 km., and from San Francisco to Honolulu, 2,080 miles, or 3,340 km.—Editor.)

of the roof. Either of these antennas can be used separately for receiving multiplex, making the use of four or more receiving sets practicable.

The operating room is on an open balcony, running across the rear of the main receiving office, which occupies a store front in the center of the city. It is subject to the ordinary noises of the street, but this causes no interference with the received signals, which are ordinarily strong and clear enough to be read under all conditions.

The operating table runs across the front of the balcony and is 12 feet (3.1 m.) long, wings on each end supporting the receiving sets. Pockets are provided at each end of the table for typewriters. Two visible-writing typewriters are used for receiving, and are so located that all tuning can be done while sitting in front of them. The sending operators sit directly beside the receiving operators. Keys operating the control circuits which run to the two transmitting-stations are available from all sending and receiving positions. Telegraph sounders and resonators are provided for the sending operators. A single telephone receiver, on an extension arm, and in series with the receiving phones of the San Francisco board, is provided for the sending operator on that circuit so that he may listen to San Francisco's "breaks."

All induction from telegraph control circuits has been eliminated by the placing of condensers of one microfarad capacity across the line, and the induction from the automatic telephone call system on the main floor has been nullified by the placing of resistance-coils in the telephone ground lead.

All messages, as they are received from the distant stations, are copied directly, in duplicate, on the typewriter, and dropped into a chute (conveniently located beside the receiving operator), which deposits them on the delivery desk on the main floor. Outgoing messages are delivered on the operating table on the balcony, by a device similar to the trolley systems used in department stores. In the handling of relay messages, it is simply a matter of one operator handing them to another.

Two transmitting stations are controlled from this office and are practically duplicates of each other, in so far as power and equipment are concerned, and are three and five miles (5 and 8 km.) from the downtown receiving office. Each transmitting station is equipped with a 12 kilowatt Poulsen arc converter, altho normally less than half of this amount of power is used. Both are arranged for transmitting upon several wave lengths,

the corresponding waves at the two stations being of exactly the same length, so that in case of the failure of one the other may be immediately put on the wave length being used, and transmission resumed without delay. This permits of one station being operated singly on holidays, etc., when there is not sufficient business to warrant keeping two stations in operation.

Normally one station is used for transmitting to San Francisco on a wave length of 3,250 meters, while the other, on a wave length of 3,750 meters, is used for transmitting to San Diego, Phoenix, and the steamships equipped with Federal apparatus, on the Pacific (which constitute a fair sized fleet). Other wave lengths are available for a second sending circuit to San Francisco when business warrants. All the company's sending stations are equipped for sending and receiving on the sustained waves of the Poulsen system. In sending on 600 meters, a "chopper" is used in series with the antenna. This furnishes a very pure musical note, at a frequency of about 500 cycles. The damped wave receivers can be instantly cut in for short wave work. Each station is in charge of a first-grade commercial radio operator. These operators are available for duty at the receiving offices when required, and the operators at the receiving offices are available for duty as arc operators at the transmitting stations. Considerable flexibility in working staff is obtained thru this arrangement.

San Diego and Phoenix normally transmit to Los Angeles on the same wave length, so that no tuning is required to hear either of these stations at any time. In addition, these stations have an auxiliary wave length exactly corresponding to that used by San Francisco, so that they can call Los Angeles on the San Francisco receiving set, or they can use that as a regular wave length when only one man is on duty at the Los Angeles office. During the busy part of the day three, and sometimes four, men are on duty at the Los Angeles office.

San Francisco Office

The receiving office at San Francisco is located on the eighth floor of a twenty-story building, and is provided with two receiving antennas of the same type as those at Los Angeles. One of these antennas is used for receiving Honolulu and the ships and the other is used for receiving Los Angeles and Portland simultaneously, providing quadruplex receiving. Two transmitting stations are controlled from this office one of these being a duplicate of the two Los Angeles stations; and the other at

South San Francisco, has two transmitters and two antennas, providing triplex sending.

Honolulu, as well as Los Angeles and Portland, is copied directly on a typewriter at the San Francisco receiving office and the copies as they are received are dispatched to the delivery department on the first floor thru a pneumatic tube.

At the San Francisco receiving office five or six men are on duty at all times during the day, working side by side, and handling heavy traffic, without interference. Moreover, the San Francisco office is in the center of the interference zone of the Pacific Coast, there being five high power stations in the immediate vicinity.

The work with Honolulu has been carried on for about five years and has been very satisfactory. A regular commercial service has been maintained and most of the news service carried by the Honolulu papers has been handled by the Federal Telegraph Company. The Honolulu end of this circuit will be duplexed in the near future to take care of increasing business.

A fast duplex service is maintained with Portland, the second antenna at the South San Francisco transmitting station being used for this purpose. It was found necessary at this station to use two antennas very nearly at right angles to each other, and two arcs for transmitting on two wave lengths simultaneously. There is no noticeable interference between transmitters and receivers at any of the stations.

Only one man is required at San Diego and Phoenix, respectively. This man acts as operator and manager. He is always able, when receiving from Los Angeles, to "break" that station when it is necessary to answer the telephone, wait on a customer, or attend to other duties. The stations in each case are located near the center of the cities they serve.

In operation of all circuits the methods of the wire companies have been followed to a great extent. Ex-wire operators are employed quite generally, and the operating officials are also ex-wire men. In the operation of the duplex circuits, the two stations are tuned in at the opening hour in the morning, after which there is no stop for tuning or calling. All business as it is filed is transmitted immediately. When it is necessary to "break," the sender is requested to re-transmit all material after the last word received, so that the receiver is able to fill in the missing words before proceeding, and to turn out a complete message ready for immediate delivery.

For transmitting purposes, the sending machines described

previously in this paper are used with few exceptions, and the speed of the circuits compares favorably with that of the fast wire circuits.

A simple form of amplifier for copying signals without the necessity for wearing head telephones has proven quite successful at Los Angeles in copying San Francisco in regular commercial work.

The Federal Telegraph Company is in active competition with the wire telegraph companies on the Pacific Coast, furnishing a fast reliable telegraph service thruout the entire year. Business is increasing daily and improvements are contemplated which will still further improve the service and the scope of the work. The circuits operated and the approximate distances covered are shown in Figure 1. A table showing the wave lengths used simultaneously without interference is shown in Table 1. This table does not show the actual degree of close tuning possible as these waves were necessarily arranged to prevent interference with other than Federal Telegraph Company stations.

TABLE 1

SIMULTANEOUS TRANSMISSION AND RECEPTION

	Transmits λ (Meters)	Receives λ (Meters)	Distance in Miles Between Transmitters and Receivers	Transmitter Power in Killowatts
San Francisco....	3,500	3,250	5	6
	10,000	10,500	7†	40
	7,500	8,000	7†	30
Los Angeles......	3,250	3,500	5	6
	3,750	2,750	3	6
Portland.........	8,000	7,500	5	30
Honolulu........	10,500	10,000	–*	40
San Diego.......	2,750	3,750	–*	4
Phoenix.........	2,750	3,750	–*	10

* Does not receive while sending.

† Same transmitting station using two transmitting sets, and with two antennas employed.

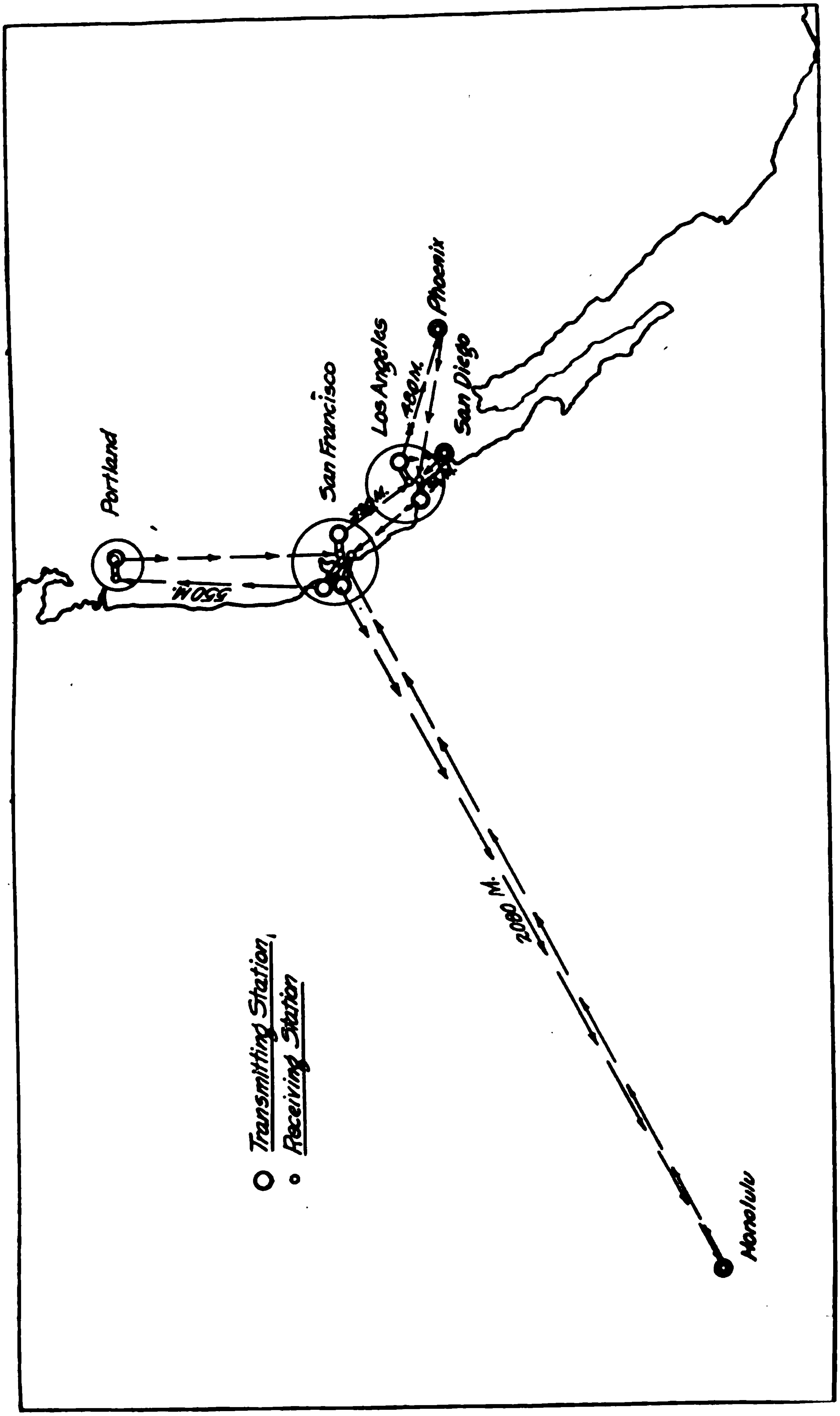
Portland
San Francisco
Los Angeles
San Diego
Phoenix
Honolulu
550 M.
2080 M.
Transmitting Station
Receiving Station

FIGURE 1

The chief factor of success has been the ability to furnish clear, reliable signals for receiving, with a minimum expenditure of transmitted power, and with consequent low cost of operation. Good working conditions for employees, and the introduction of the latest time-saving features have also proven big factors in efficiency. Flexibility in power control and wave length regulation and the sharp tuning have assisted materially.

In the work of these stations, the matter of interference from extraneous noises has been shown to be more of a psychological than a material difficulty. An operator, trained to receive with outside disturbances present, listens only to the signals he wishes to copy, regardless of the fact that other noises around him may be considerably louder. Soundproof compartments for receiving have been demonstrated to be entirely unnecessary in this work.

The great value of the duplex system over the old style simplex station for high speed, direct work, has been the most striking feature. With the results of this work in view, the erection of simplex stations in the future, to handle any considerable amount of business, would seem to be an economic waste. In comparison: a duplex radio circuit would seem to excel a simplex non-break installation in the proportion of about 3 to 1.

SUMMARY: After considering some of the obstacles in the way of successful competition of overland radio service versus wire service, the author treats the mode of overcoming these difficulties. He recommends also radio duplex circuits; reception with loud-speaking receivers and amplifiers; trunk and way circuits from large radio centers of traffic; and relaying stations. The organization and operation of the Pacific coast chain of duplex radio stations of the Federal Telegraph Company is hen described in detail.

A SPECIAL TYPE OF QUENCHED SPARK RADIO TRANSMITTER*

By

D. Galen McCaa

(Research Engineer, John Firth, New York City)

The transmitter to be described consists partly of a special arrangement of the conductors forming the antenna; the structure presenting other circuits in conjunction with the usual open radiating circuit.

The entire transmitter is represented in Figure 1.

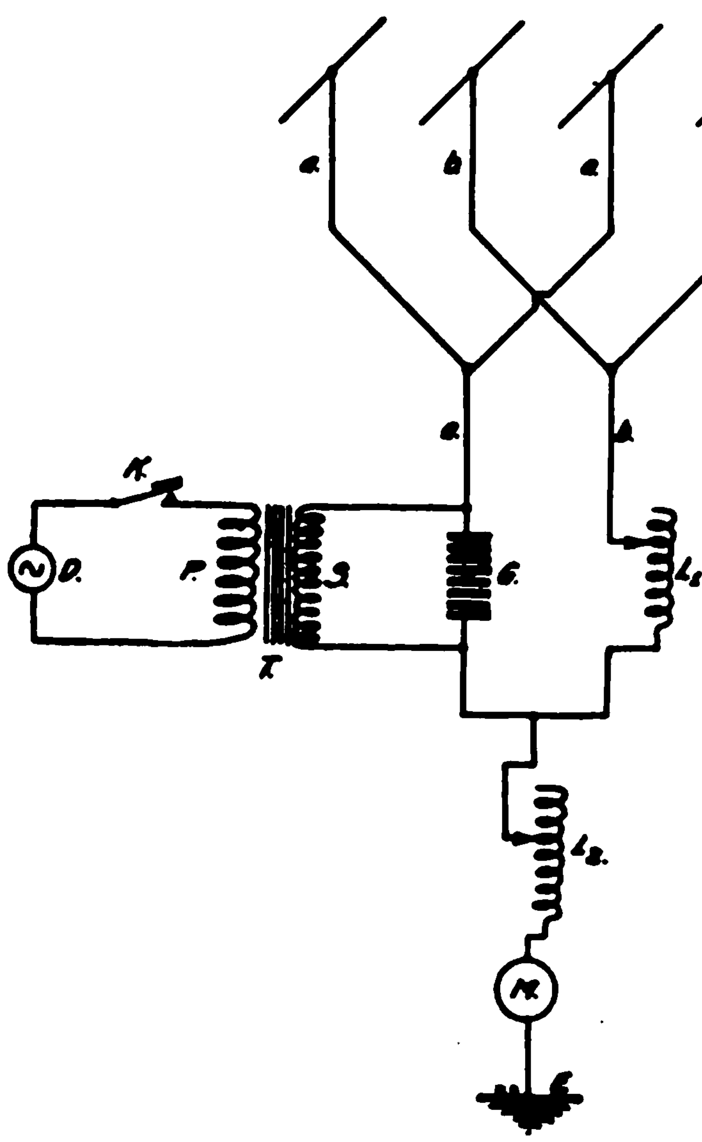

Figure 1

A number of horizontal and vertical wires, *A*, forming a "T" or an "L" type antenna, have alternating between and insulated from them a like number of wires, *B;* the wires of each

* Received by the Editor, October 15, 1918.

group being connected in parallel, and connections from each group being led to the instruments; which consist of a quenched gap, G, two inductance coils L_1 and L_2, a transformer T, a key K, and an alternator D. The earth connection is made thru the inductance L_2 and ammeter M.

Three oscillatory circuits are formed by this structure—first: the conductors A, gap G, inductance L_2, and earth E, which form an open circuit oscillator; second: the conductors A, gap G, inductance L_1, and the conductors B, which form a partially closed circuit oscillator; third: the conductors B, inductances L_1 and L_2, and the earth E, which form an open circuit oscillator.

The gap is common to the circuits A, G, L_2, E and A, G, L_1. B. These circuits are, therefore, the primary oscillatory systems and supply energy to the structure B, L_1, L_2, E which is the radiating system.

The capacity available is determined by the capacity of the conductors A to earth, which is markedly increased by the proximity of the earthed conductors B, and by the capacity found between the group of conductors A and the group B. These two capacities are charged in parallel. It may be noted here that the conductors B serve the following purposes: first, augment the capacity of the conductors A; second, form a primary oscillator in conjunction with the conductors A, and third, become the capacity of an open circuit oscillator.

The primary circuit A, G, L_2, E, is coupled to the radiating circuit B, L_1, L_2, E by the inductance L_2. This coupling is varied by the inductance L_1. The primary circuit A, G, L_1, B is coupled to the radiating circuit B, L_1, L_2, E by the inductance L_1. This coupling is varied by the inductance L_2. The circuits are closely coupled, for in addition to the electromagnetic couplings, a capacitive coupling exists between the conductors A and B.

Owing to the antenna structure employed and the close couplings presented it is imperative that the primary discharges be highly damped. With a single stationary zinc spark gap, there will be found two sets of coupling oscillations in the circuit B, L_1, L_2, E, and the structure A remains a part of the radiating system during the primary discharge, and it will be carrying current opposite in direction to the structure B. The effect will be similar to that found in a loop antenna.

The close and fixed couplings do not lend themselves readily to the use of true quenched gap effects, where the quenching is partly determined by the reaction of the secondary current.

In order that there shall be no effectual radiation from the two primary circuits, the energy of the primary oscillations must be quickly transferred to the secondary system and the gap must become an open circuit after the first few oscillations. This result is obtained by employing a very high group frequency of highly damped primary discharges.

With the usual closed circuit primary oscillator consisting of a quenched gap, condenser, and inductance, it is possible to obtain group frequencies of forty-thousand discharges per second, the number depending upon the design of the oscillator and its relation to the supply system. The damping of the oscillations in such a system is determined by the supply current, the design of the gap, and the constants of the primary oscillator. The higher the group frequency the greater will be the decrement of each primary discharge; and with the group frequencies actually employed a type of impulse excitation results.

Discharges of this type will occur in a primary oscillatory circuit when it is not coupled to a secondary circuit. The reaction of a secondary system is, therefore, not required as in true quenched gap operation.

Figure 2 is a photographic record of the discharges in an oscillator designed to produce high group frequencies, and shows the discharges that occur during the time period of one alternation; the discharge frequency is approximately thirty-five-thousand per second along the crest of the alternating current wave.

FIGURE 2

Measuring the logarithmic decrem
in the primary system will not indicat
in each discharge as the gap decremen
determined by oscillograph records).

While ideal impulse excitation—a
current—may not be present, a very
attained as is indicated by the fact t
wave length curves are almost flat.

With primary discharges as described, the gap is quickly rendered an open circuit, the primary systems A, G, L_2, E and A, G, L_1, B cease to exist as such, and the secondary system B, L_1, L_2, E is left free to oscillate at its own natural period and with a decrement determined solely by its constants.

The secondary oscillations, when high group frequencies are employed, are continuous tho not undamped. This is consequent upon the decrement of the antenna oscillations and the group frequencies. With a low decrement and high group frequency, the antenna will receive a second impulse before its oscillations have materially declined. It is probable that the electromotive forces impressed on the gap by the secondary current "trigger" it off when adding to the supply electromotive force, and the resultant discharge in the primary is then in phase with the antenna oscillations.

The adjustments of the inductances L_1 and L_2 for maximum radiation, as indicated in the ammeter M, are not critical owing to the type of impulse excitation employed; and, while a maximum can be found, it exists thruout broad adjustments of the inductances. If, with the inductances adjusted for maximum radiation, the system is analyzed by wave length measurements, leaving the inductance values fixed, the primary circuit A, G, L_2, E will present a shorter wave length and the primary circuit A, G, L_1, B will present a longer wave length than the wave radiated by B, L_1, L_2, E. The wave radiated is determined by the capacity of the conductors B to earth and the inductance values of L_1 and L_2; on inserting a spark gap in this structure and exciting it as a primary oscillator, substantially the same wave length is found.

The dissonance between the primary systems and the radiating antenna has been found to be as high as 20 per cent, the exact value varying with the group frequency employed. This is far in excess of the 2 per cent dissonance found in true quenched gap action.

Figure 3 presents the resonance curves obtained in one case.

The curve A was read from L_2 when the primary circuit A, G, L_2, E was excited alone, the conductors B being grounded independently. The curve C was read from L_1 when the primary circuit A, G, L_1, B was excited alone, the ground lead and L_2 being removed. The curve B was read from the inductance

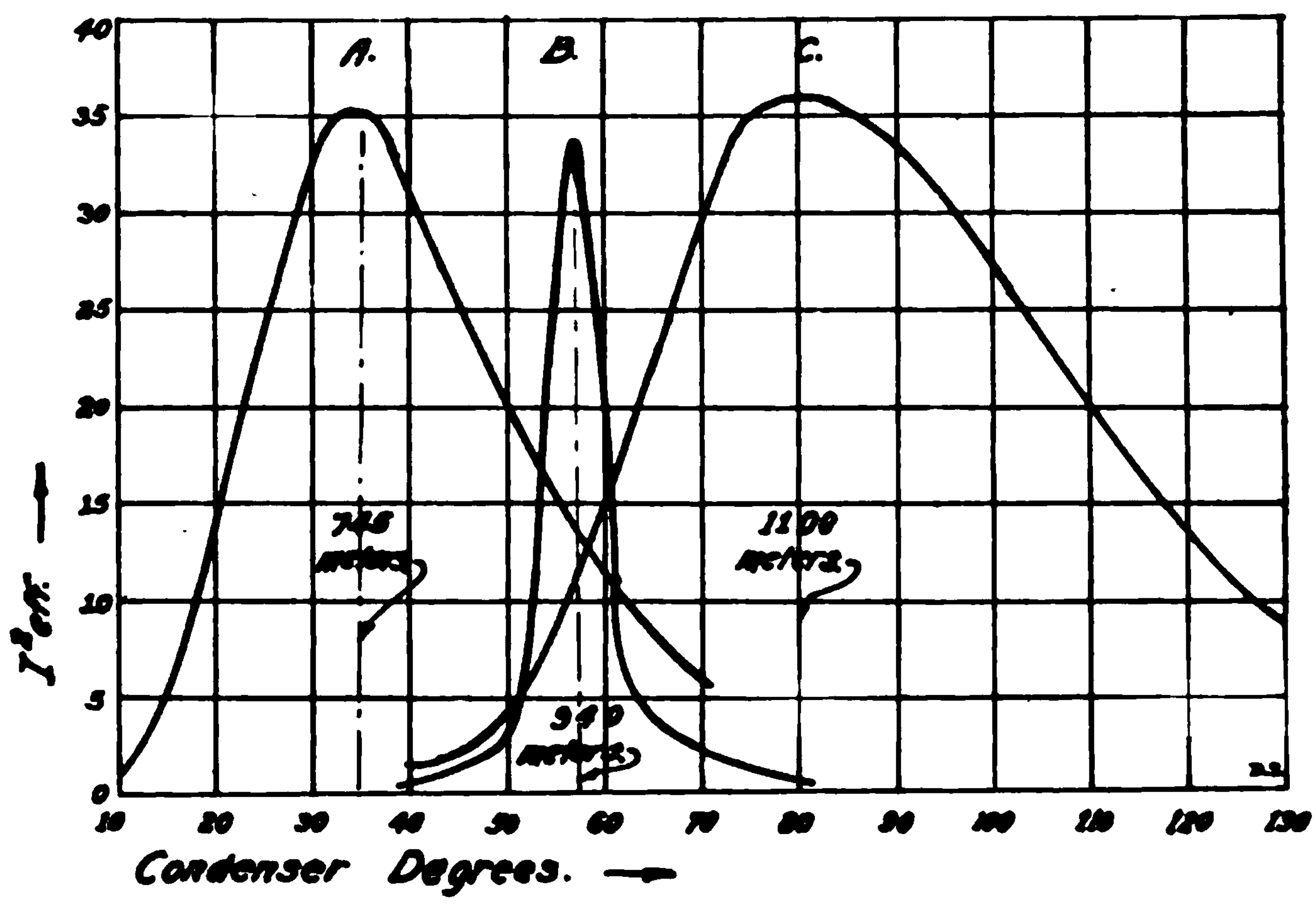

FIGURE 3

L_2 when both primary circuits were exciting the circuit B, L_1, L_2, E; and is the wave radiated. The wave lengths in this case were 745 meters for A, 940 meters for B, and 1,100 for C. It may be of interest to note that the curve B shows no evidence of the oscillations of the primary circuit A, G, L_2, E when read from L_2, and a curve read from L_1 shows the same symmetry, there being no evidence of the oscillations of the primary circuit A, G, L_1, B.

By placing a single turn coil in the conductor A immediately above the gap, both the primary wave lengths can be found. They, therefore, exist in L_1 and L_2, but are overshadowed by the more powerful oscillations of the antenna.

It will be noted that the ammeter is so placed as to be not only in the radiating antenna, but also in the primary circuit A, G, L_2, E. If the ammeter is placed immediately below the inductance L_1, the reading will be the same as if placed immediately above L_2. The values of the primary currents are usually different, and if the primary value is added to the true antenna current, different readings would be expected, dependent

on whether the ammeter is placed in one primary or in the other.

The dissonance found in this system is not consequent upon the arrangement of the circuits, but the circuits are operative *because of* the dissonance.

Between a single primary oscillator made up of a condenser, quenched gap, and inductance so designed as to produce high group frequency, and an antenna, the same dissonance effect has been found. That the structure is operative owing to the dissonance will be apparent from a consideration of Figure 4.

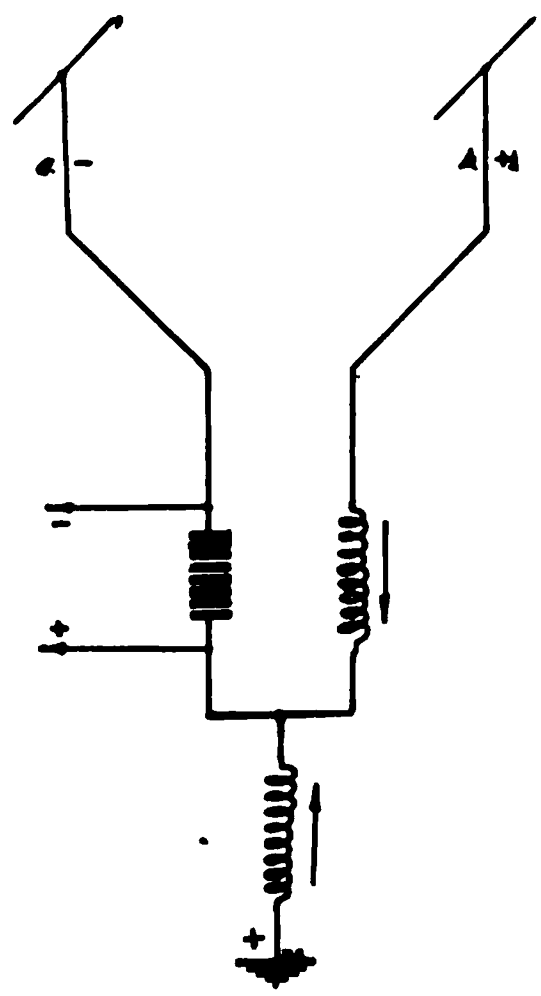

FIGURE 4

Consider the system charged as indicated by the symbols. It will be apparent that the first alternation of the discharge will flow thru the inductance as indicated by the arrows. The magnetic fields established in these auto transformers will be opposed to each other and if the currents were in phase and of equal amplitude the final result in the antennas would be zero. Cases have been observed where the three circuits were of the same period and the system was practically inoperative. With a large difference between the frequencies of the two primary systems, there is less opposition between the two auto transformers, and owing to the impulse excitation, the primaries are capable of exciting an antenna not in resonance with them, but one the period of which lies between the periods of the primaries.

Attempts have been made to reduce the phenomena in an artificial antenna or phantom circuit, but the results were not the same as found in the aerial structure.

SUMMARY: A quenched spark transmitter is so arranged that the capacity in the highly damped primary circuit is that between a special extra antenna and ground, and the primary and secondary circuits are partly inductively coupled thru a common inductance in the ground lead and partly capacitively coupled by the capacity between the special antenna and the usual secondary or radiating antenna.

Quenching effects and normal mono-wave radiation are secured. Experiments are described and an oscillogram shown whereby the group frequency and radiation characteristics are indicated.

ON THE MULTI-SECTION QUENCHED GAP*

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It is well known that very short spark gaps (about 0.2 mm or 0.008 inches in length) possess the valuable property of causing quenching of a spark discharge. On the other hand, such gaps have a low breakdown voltage, of the order of 800 or 900 volts. This defect as regards ease of manipulation can be readily overcome by connecting a number of short gap sections in series. We thus obtain a multi-section quenched gap possessing the property of quenching because of the characteristic of the separate sections, and yet having a breakdown voltage which can be varied within wide limits by a suitable choice of the number of sections employed.

It is usually agreed that the total voltage required for a multi-section quenched gap is directly proportional to the number of sections, and that, therefore, by increasing the number of gaps, the discharge potential may be increased to any desired extent. In the following discussion, it will be shown that such an assumption is not correct, and that there exists an upper limit of voltage which cannot be exceeded by further addition of new gap sections to the series. The cause of this limitation lies in the uneven distribution of potential or potential gradient along the series of gaps.

Let us consider, for example, a quenched spark gap of the Telefunken type, consisting of fairly large, circular, metallic plates separated by thin insulating rings, and with a voltage

* Received by the Editor, September 5, 1918.

across the total gap of $V=P_s-P_o$, before the gap has broken down, as indicated in Figure 1.

FIGURE 1

If we neglect the induction current due to imperfect insulation, we know that the total current between the plates due exclusively to variation of dielectric displacement is

$$I=\frac{dD}{dt}.$$

It is clear that the total electric flux starting say from plate $(n-1)$ does not exclusively enter the following plate, (n). Some of the lines of induction will pass directly to the oppositely charged end of the series of gaps, or will pass thru neighboring conductors to the earth, and so on. There is thus a certain leakage of the dielectric displacement current and if we denote the current entering the plate n, and, therefore, starting from it, by I_n, and the leakage current from this plate by i_n we will have for current entering the plate $n+1$, the value

$$I_{n+1}=I_n-i_n \tag{1}$$

This current can be expressed in terms of the voltage between the plates as follows:

$$I_{n+1}=(P_{n+1}-P_n)\omega C \tag{2}$$

$$I_n=(P_n-P_{n-1})\,\omega C \tag{3}$$

where ω signifies the angular velocity or $2\pi f$, C the capacity of the condenser formed by the two adjacent plates, and P_{n+1}, P_n, P_{n-1} . . . . , the potential of the corresponding plates.

We can write a similar expression for the current I_n:

$$i_n=(P_n-P_o)\,\omega c$$

or putting $P_o=0$,

$$i_n=\omega c P_n, \tag{4}$$

where by c we understand the capacity of the plate with respect to the earth, to leads, and all other neighboring conductors except the next plates.

From (1), considering (2), (3), and (4), we have

$$\frac{c}{C} P_n = P_{n+1} - 2 P_n + P_{n-1} \tag{5}$$

The solution of this equation can be put in the form of

$$P_n = A \varepsilon^{an} \tag{6}$$

For P_{n+1} and P_{n-1} we then obtain

$$P_{n+1} = A \varepsilon^{a(n+1)} = \varepsilon^{a} A \varepsilon^{an} \tag{6'}$$

$$P_{n-1} = A \varepsilon^{a(n-1)} = \varepsilon^{-a} A \varepsilon^{an} \tag{6''}$$

By substitution we get from (5):

$$\frac{c}{C} = \varepsilon^{a} - 2 + \varepsilon^{-a} = (\varepsilon^{\frac{a}{2}} - \varepsilon^{-\frac{a}{2}})^2 = \left(2 \sinh \frac{a}{2}\right)^2 \tag{7'}$$

Putting $\frac{c}{C} = k$, we obtain

$$\sinh \frac{a}{2} = \tfrac{1}{2} \sqrt{k} \tag{7}$$

As the solution of (5) can also be put into the form $P_n = B \varepsilon^{-an}$, we may write the more general expression for the potential of the n-th plate, P_n, namely:

$$P_n = A \varepsilon^{an} + B \varepsilon^{-an}. \tag{8}$$

The constants A and B depend upon the values of P_o and P_z, these latter being the values of the potentials at the ends of the series of gaps.

In radio practice, two different cases may arise (of which the first is):

$$P_o = 0; \; P_z = V_{max},$$

V_{max} being the voltage produced by the transformer when the spark circuit is directly coupled to the antenna; and (for the second case):

$$P_o = -\frac{V_{max}}{2}; \; P_z = +\frac{V_{max}}{2},$$

when the coupling to the antenna is inductive, since in this case the neutral point of the transformer is usually connected to ground.

In the first case we have for the first of the series of gaps, where

$$n = 0,$$

$$A + B = 0;$$

and therefore

$$A = -B,$$

while at the end of the series of gaps, where $n = z$,

$$A\,\varepsilon^{az} + B\,\varepsilon^{-az} = A\left(\varepsilon^{az} - \varepsilon^{-az}\right) = V_{max}$$

and therefore

$$A = \frac{V_{max}}{\varepsilon^{az} - \varepsilon^{-az}} = \frac{V_{max}}{2\,sinh\,a\,z}.$$

Substituting these values of the expression for the potential of the n-th plate, we get

$$P_n = V_{max}\frac{sinh\,a\,n}{sinh\,a\,z} \tag{9}$$

which expresses the law governing the distribution of potential along the series of plates of the gap.

This potential distribution is shown graphically in Figure 2.

FIGURE 2

From this, we note that the voltage applied to each gap of the series is by no means the same, but increases as n, the number of gaps, increases. If we denote the voltage between the plate n and plate $n-1$ by v_n, we have

$$V_n = P_n - P_{n-1} = \frac{V_{max}}{\sinh a z}(\sinh a n - \sinh a (n-1))$$

and since

$$\sinh a n - \sinh a (n-1) = 2 \sinh \frac{a}{2} \cosh \frac{a(2n-1)}{2},$$

we obtain directly

$$v_n = V_{max} \frac{2 \sinh \frac{a}{2}}{\sinh a z} \cosh \frac{a(2n-1)}{2}. \tag{10}$$

This expression can be put into another form which permits us to draw certain interesting conclusions. The form referred to is

$$v_n = V_{max} \frac{2\left(\varepsilon^{\frac{a}{2}} - \varepsilon^{-\frac{a}{2}}\right)\left(\varepsilon^{an-\frac{a}{2}} + \varepsilon^{-an+\frac{a}{2}}\right)}{\varepsilon^{az} - \varepsilon^{-az}}$$

$$= V_{max}\left(1 - \varepsilon^{-a}\right) \frac{\varepsilon^{an-\frac{a}{2}} + \varepsilon^{-an+\frac{a}{2}}}{\varepsilon^{az-\frac{a}{2}} - \varepsilon^{-az-\frac{a}{2}}}$$

As the number of plates n increases, this expression approaches the limit

$$v_n = V_{max}\left(1 - \varepsilon^{-a}\right) \tag{11'}$$

or

$$V_{max} = \frac{v_n}{1 - \varepsilon^{-a}} \tag{11}$$

The breakdown voltage of a given gas being fixed, we see from this that the total voltage V_{max} applied to a series of gaps cannot be made to exceed the value determined by equation (11) above.

The increase in breakdown voltage of a multi-section quenched gap with the increase in the number of sections is shown in Figure 3 for the values of $k = \frac{c}{C} = 0$; 0.000625; 0.00125, 0.0025; 0.005; and 0.01. It is evident that the upper limit of spark voltage is reached for $k = 0.01$ when $n = 24$, for $k = 0.005$ when $n = 34$, and so on. Furthermore, this limit for $k = 0.01$ is about 10 times the breakdown voltage of the short gap section; while for $k = 0.005$ it is 14.6 times the same voltage, and so on.

The effect of the flux leakage on the highest obtainable gap voltage of a multi-section discharger is more fully shown by the curve of Figure 4, the abscissas being the ratio $k=\frac{c}{C}$, and the ordinates showing the greatest attainable multiple of the breakdown voltage of a short gap for a multi-section gap consisting of such short gaps connected in series.

FIGURE 3

On considering the second of the above-mentioned possibilities; namely, that of inductive coupling of the spark circuit to the antenna circuit, we have

at $$n=0$$

$$-\frac{V_{max}}{2}=A+B$$

and at

$$n = z$$

$$+\frac{V_{max}}{2} = A\,\varepsilon^{az} + B\,\varepsilon^{-az}$$

FIGURE 4

Therefore

$$A = \frac{V_{max}}{2} \cdot \frac{1+\varepsilon^{-az}}{2\sinh a z}$$

$$B = -\frac{V_{max}}{2} \cdot \frac{1+\varepsilon^{-az}}{2\sinh a z}$$

and the expression for the potential of the n-th plate takes the form

$$P_n = \frac{V_{max}}{2} \cdot \frac{\sinh \frac{a}{2}(2n-z)}{\sinh \frac{a z}{2}} \tag{12}$$

This distribution of potential is shown by the curves of Figure 5, these curves being drawn for the same values of k, as in Figure 2; namely, $k=0.04$; 0.03; 0.02; 0.01; and 0. By comparison with Figure 2, we see that the distribution or gradient potential is much more uniform, and that the middle gap sections in this case have less stress on them, the excess voltages being equal at each end of thé series of gaps.

FIGURE 5

For the voltage applied to the gap section between plate n and plate $n-1$, we obtain

$$v_n = P_n - P_{n-1} = V_{max} \frac{\sinh \frac{a}{2}}{\sinh \frac{a z}{2}} \cdot \cosh \left[\frac{a}{2} (2n - z - 1) \right] \quad (13)$$

The voltage of the last gap of the inner series will be

$$v_z = V_{max} \frac{\sinh \frac{a}{2}}{\sinh \frac{a z}{2}} \cdot \cosh \frac{a}{2} (z-1).$$

Putting this under the form

$$v_z = V_{max} \frac{(1-\varepsilon^{-a})}{2} \cdot \frac{\varepsilon^{\frac{a}{2}(z-1)} + \varepsilon^{-\frac{a}{2}(z-1)}}{\varepsilon^{\frac{a}{2}(z-1)} - \varepsilon^{-\frac{a}{2}(z+1)}},$$

we see that with increasing z, this tends toward the limit

$$v_z = V_{max} \cdot \frac{1-\varepsilon^{-a}}{2} \tag{14'}$$

From this we obtain

$$V_{max} = \frac{2\,v_z}{1-\varepsilon^{-a}} \tag{14}$$

Consequently, for this case of the inductive coupling of the spark and antenna circuits, just as in the preceding case for direct coupling, there exists an upper limit to the discharge voltage of the entire series of gaps. When we are given the breakdown voltage of the individual section and the "flux leakage," this limiting value is twice as high in the case of the inductive coupling as in the case of direct coupling.

SUMMARY: The authors consider the relation between the breakdown voltage of a series of quenched gap sections and that of a single section. Because of electric flux leakage from each plate to nearby plates and neighboring conductors, the relation of direct proportionality does not hold. The breakdown voltage of a number of gaps of given length can not be made to exceed a limiting value, given in the paper. The limiting value in question is shown graphically for various values of flux leakage and breakdown voltage of gap.

When spark circuit and antenna circuit are coupled magnetically, the available limiting breakdown voltage is twice that for direct coupling with one side of the high voltage transformer grounded.

A STUDY OF ELECTROSTATICALLY COUPLED CIRCUITS*

By

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With a view to justifying an extended investigation of electrostatic coupling—if such a justification be needed—let us consider one application which this kind of coupling might have in radio communication. This is the problem of producing harmonic oscillations in an antenna. The advantage of such an arrangement is apparent to an experimenter who desires to use a large antenna for receiving signals, and yet who, on account of the law or for other reasons, must use in transmitting, a wave length which is but a fraction of the natural, or fundamental, wave length of the antenna circuit. If his antenna circuit could be made to oscillate with a wave length which is the first harmonic of the fundamental, there would be emitted a wave which would have only one-third the length of the fundamental. If this were not short enough for the purposes, the second harmonic might be used.

Just before private radio communication was prohibited, I was able to produce these harmonics in the antenna which had, with the rest of its circuit, a fundamental wave length of 665 meters. I had time to try only magnetic coupling between circuits. When the primary circuit of the transmitting set was tuned to the same wave length as the fundamental of the antenna circuit, it was found that the antenna circuit oscillated fundamentally. (This statement is true, assuming that the coupling between circuits is not close enough to allow the secondary to react on the primary, thereby giving two waves, one above and one below the one otherwise expected.) When the primary circuit was tuned to a wave length one-third that of the fundamental, this first harmonic was obtained in the antenna with no trace of the fundamental or any other wave length. By reducing the wave length of the primary to one-fifth that of the fundamental, the second harmonic was obtained in the antenna.

*Received by the Editor, July 26, 1918.

For the fundamental, the primary circuit probably has only enough inductance to give the needed coupling to the antenna circuit, and all the capacity possible, consistent with the wave length, to keep the energy of the circuit at a high value, having a fixed potential available. Now if we wish to produce the first harmonic, we must reduce the product of inductance and capacity to one-ninth its value for the fundamental. But since our inductance is already no more than we need for our magnetic coupling, it must be the capacity, and consequently the energy of our system which is reduced, since $W=\frac{1}{2}CV^2$.

Electrostatic coupling between the circuits offers a solution to this problem, for we can then reduce the inductance without affecting either the coupling or the energy of the circuit.

So it was that the foregoing work seemed to lead logically to a study of the characteristics of electrostatic coupling. Mr. Laurens E. Whittemore of the Physics Department of the University of Kansas was just beginning such a study, and so we carried on the work together.

Our purpose in this research was first to investigate the mathematical theory of electrostatically coupled circuits and to test experimentally the truth of the conclusions drawn, and secondly to study by means of the Braun tube and sustained oscillations the relations existing between the variables in the electrostatically coupled circuits using various values for the coefficient of coupling.

E. Bellini[1] has worked out the mathematical theory of the general case of electrostatically coupled circuits, such as in Figure 1, by solving the differential equations which may be set up for the circuits from Kirchoff's laws. Mr. Whittemore took the easier way, and solved the equations set up in complex notation form. I will merely outline Mr. Whittemore's work.

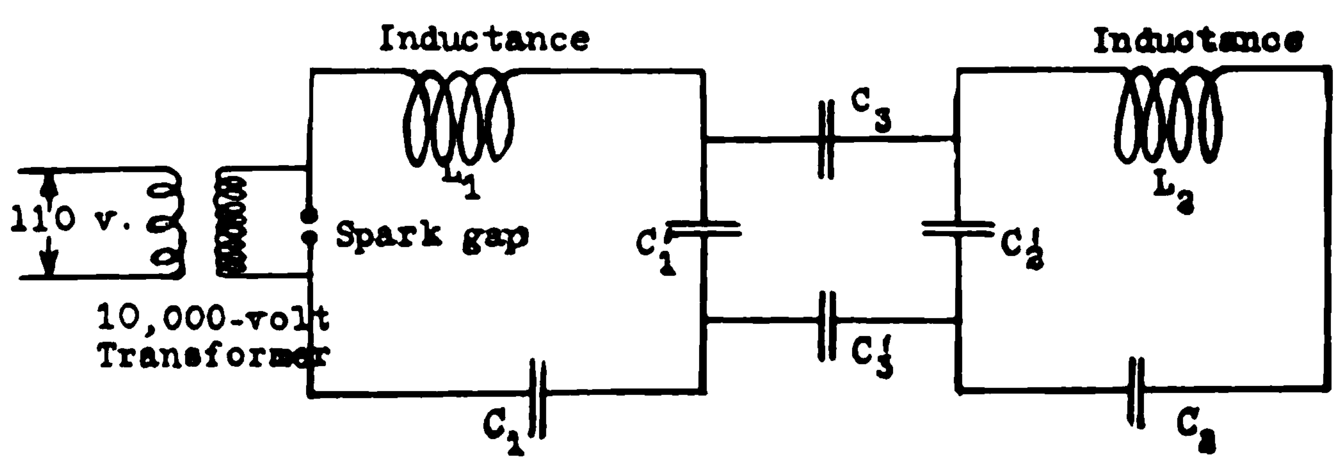

FIGURE 1—Electrostatically Coupled Circuits

[1] "La Lumière Electrique," volume 32, page 241, 1916.

The equations for potential drops and currents in the primary, secondary, and intermediate circuits are the following:

$$i_1\left(j\,p_1 L_1+\frac{1}{j\,p_1 C_1}\right)+\frac{i_1'}{j\,p_1 C_1'}=0$$

$$i_2\left(j\,p_2 L_2+\frac{1}{j\,p_2 C_2}\right)+\frac{i_2'}{j\,p_2 C_2'}=0$$

$$\frac{i_1'}{j\,p_1 C_1'}+\frac{i_3}{j\,p_3 C_3}+\frac{i_3}{j\,p_3 C_3'}+\frac{i_2'}{j\,p_2 C_2'}=0$$

$$i_1'=i_1+i_3$$

$$i_2'=i_2+i_3$$

Assuming that $p_1=p_2=p_3$, equating the determinant to zero, and solving for p^2

$$p'^2=\frac{1}{2}\left(\frac{1}{L_1G_1}+\frac{1}{L_2G_2}\right)-\sqrt{\frac{1}{4}\left(\frac{1}{L_1G_1}-\frac{1}{L_2G_2}\right)^2+\frac{k_e^2}{L_1G_1L_2G_2}}$$

$$p''^2=\frac{1}{2}\left(\frac{1}{L_1G_1}+\frac{1}{L_2G_2}\right)+\sqrt{\frac{1}{4}\left(\frac{1}{L_1G_1}-\frac{1}{L_2G_2}\right)^2+\frac{k_e^2}{L_1G_1L_2G_2}}$$

where
$$\frac{1}{G_1}=\frac{1}{C_1}+\frac{1}{C_1'}-\frac{C_t}{C_1'^2}$$

and
$$\frac{1}{G_2}=\frac{1}{C_2}+\frac{1}{C_2'}-\frac{C_t}{C_2'^2}$$

$k_e^2=\left(\frac{C_t}{C_1'C_2'}\right)^2G_1G_2$ where k_e is the coefficient of coupling.

$$\frac{1}{C_t}=\frac{1}{C_1'}+\frac{1}{C_2'}+\frac{1}{C_3}+\frac{1}{C_3'}$$

Taking the special case where the frequencies of the two circuits are the same *before* coupling

$$\frac{1}{L_1}\left(\frac{1}{C_1}+\frac{1}{C_1'}\right)=\frac{1}{L_2}\left(\frac{1}{C_2}+\frac{1}{C_2'}\right)=4\,\pi^2 n^2$$

where n is the natural frequency of each circuit; from which we get

$$p'=2\,\pi\,n$$

$$p''=4\,\pi^2 n^2-\frac{C_t}{L_1\,C_2'^2}-\frac{C_t}{L_2\,C_2'^2}$$

$\frac{\lambda'}{\lambda''}=\sqrt{\frac{1-k_e}{1+k_e}}$ where λ' and λ'' are wave lengths.

When the frequencies of the two circuits are the same *after*

coupling, that is, when each circuit is tuned to the same wave length with the intermediary coupling condensers connected,

$$L_1 G_1 = L_2 G_2;$$

and we get

$$p' = \sqrt{\frac{1-k_e}{L_1 G_1}}$$

$$p'' = \sqrt{\frac{1+k_e}{L_1 G_1}}$$

$$n' = n\sqrt{1+k_e} \quad \text{or} \quad \lambda' = \frac{\lambda}{\sqrt{1+k_e}}$$

$$n'' = n\sqrt{1-k_e} \quad \text{or} \quad \lambda'' = \frac{\lambda}{\sqrt{1-k_e}}$$

In both cases, the shorter wave corresponds to the natural frequency of one of the single circuits without the coupling condenser.

To comply with the condition that each circuit have the same frequency before coupling, we excited each circuit separately (when the coupling capacities were not connected), by a spark gap and transformer, and tuned each to the desired wave length, as determined by a resonating wave meter. The circuits were then connected electrostatically and the system excited, using a spark gap in one circuit. The waves present in each circuit were determined with the wave meter.

For the other condition, we tuned each circuit when the other circuit was broken at some point other than between the coupling connections. The connections were then made and the wave lengths in each circuit determined. The theory was tested for the extreme values of coupling as well as for a number of intermediate values. We sometimes increased capacities conveniently by merely short-circuiting the condenser.

The following are the data for our observed and calculated values, from which Figure 2 is plotted:

OBSERVED					CALCULATED			
k_e	λ	λ'	λ''	$\frac{\lambda''}{\lambda'}$	λ'	λ''	k_e	$\frac{\lambda''}{\lambda'}$
.80	...	290	908	3.13	290	872	.0	1.00
.50	...	407	727	1.78			.1	1.10
.50	510	414	722	1.74	417	722	.2	1.22
.00	...	525	525	1.00			.3	1.36
.20	...	285	358	1.26			.4	1.53
.20	315	285	355	1.25			.5	1.73
.32	...	405	597	1.47	405	564	.6	2.00
.43	...	296	465	1.57			.7	2.38
.50	...	300	515	1.71	300	520	.8	3.00
.50	363	291	520	1.79			.9	4.36
.43	355	300	475	1.58	298	470	1.0	∞
.40	450	390	553	1.42	381	582		
.20	342	322	375	1.16				
.20	450	410	495	1.21				
.14	215	200	237	1.18	201	233		
.14	...	198	243	1.22	198	232		
1.00	317	225	∞	∞				
1.00	317 555	278	∞	∞				
1.00	153 555	150	∞	∞				
1.00	218	157	∞	∞				

The curve (Figure 2) gives a good comparison of observed and calculated values. We see that as our co-efficient of coupling, k_e approaches more and more closely to unity, one wave length approaches infinity under both conditions of tuning. When the circuits are in tune after coupling, while the one wave length approaches infinity, the other approaches zero. But

FIGURE 2

in the case of tuning before coupling, we have one wave always the same, while the other approaches infinite length with no energy content. This is surely an ideal state of things, since we can then transfer nearly all the energy of the primary to the secondary and yet have that energy in only one wave.

The Braun tube method of studying the arc phenomena of single circuits is not by any means new, but we are reasonably sure that the effects produced by electrostatically coupling a secondary circuit have never been investigated by this method.

Professor Simon[2] explored the "dynamic characteristic" of an alternating current arc by means of a Braun tube arranged so that the cathode ray pointer was deflected horizontally by the arc current and vertically by the potential difference across

[2] H. Th. Simon, "Phys. Zeitschr.," volume 6, page 297, 1905; volume 7, page 423, 1906.

the arc. The closed curve he obtains is of the form of Figure 3 and shows a phenomenon called "arc hysteresis." This shows very clearly how the variables are related and that the arc actually has a falling characteristic. In 1900, Mr. Duddell[3] showed that a direct current arc gave out a musical note when it was shunted by a condenser and an inductance, both of proper proportions. The most extensive and valuable study of the dynamic characteristics of the oscillating arc was made by Simon and his students.

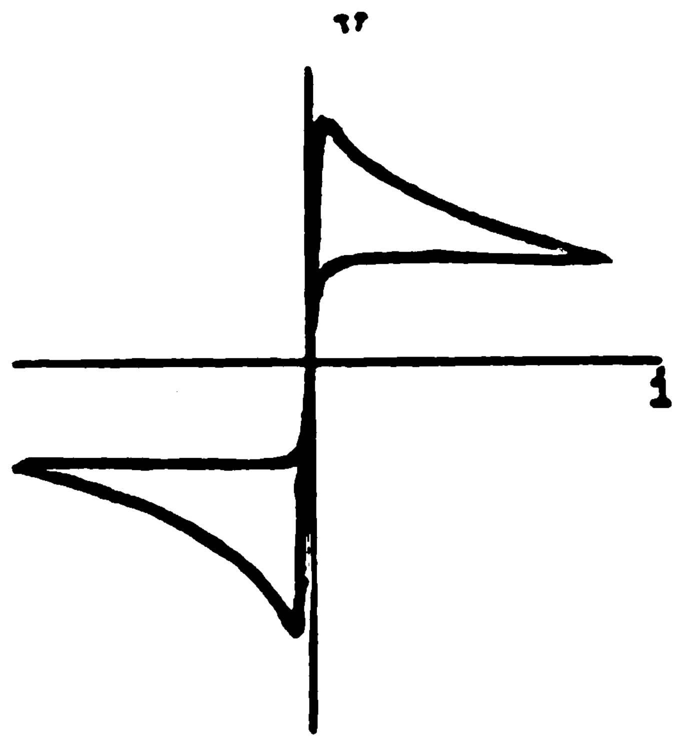

FIGURE 3

Mr. Hidetsugu Yagi[4] has investigated the reacting effect of a magnetically coupled secondary circuit on the oscillation of a carbon arc. In our experiments, we are substituting known values of electrostatic coupling for his magnetic coupling.

There are three types of oscillations which may be obtained with an arc. If there were no oscillations the current thru the arc, i_o, would be nearly constant. The condenser discharge thru the arc tends to superpose a sinusoidal current and make the current pulsating. So long as i_o is larger than the amplitude of pulsation, there is no extinction of the arc, and the oscillation is said to be of the "first type." The oscillation of this type is generally obtained in musical arcs. When the fluctuation becomes larger than i_o, there will be a period of zero current, and the arc will be extinguished for a moment. If the arc extinguishes, a constant current, i_o, will flow into the condenser and charge it up until its potential becomes sufficiently

[3] W. Duddell, "Journal I. E. E.," volume 30, page 232, 1900.
[4] Hidetsugu Yagi, "Proc. Inst. Radio Engrs.," volume 4, page 371, 1916.

high to cause the next discharge across the arc gap. This is called the "oscillation of the second type," and is most readily obtained in practice at radio frequencies, especially when there is any dissimilarity of electrode material. If the terminal potential difference, which becomes reversed at the extinction, is large enough to cause a discharge across the gap, it will light a small arc in the opposite direction. The oscillation with this reverse discharge is of the third type. The three types are diagrammatically represented in Figure 4, as taken from Mr. Yagi's paper.

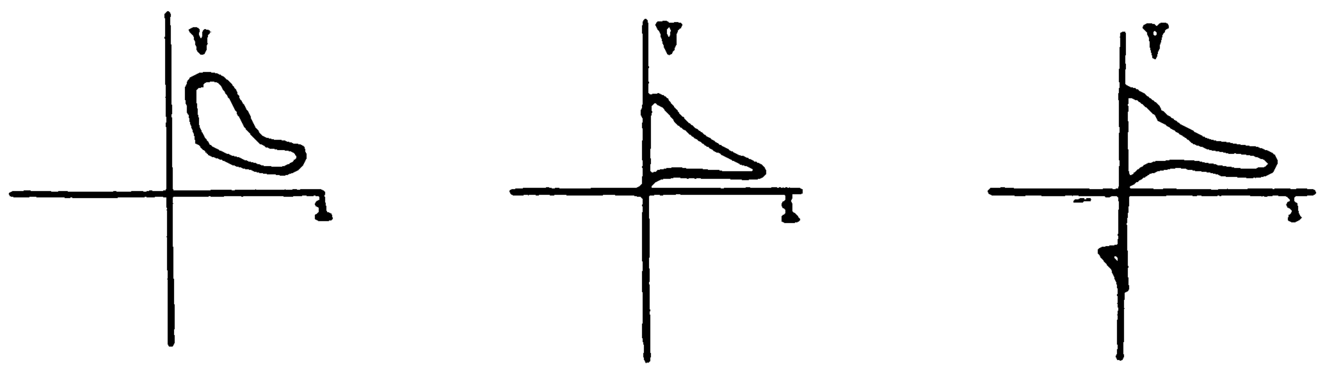

FIGURE 4

As the second type of oscillations are used in radio communication, we have used this type in our study.

In our experimental work, we studied the relations between (1) $\frac{di}{dt}$, and i in the primary, where i is current; (2) potential difference across condenser in secondary, and primary current; (3) i in secondary, and i in primary; (4) potential difference across arc, and secondary i; (5) $\frac{di_2}{dt}$, and i_2; (6) potential difference across arc, and i in the primary, which, however, was not very successful because our Braun tube was not constructed so as to give us the necessary amplitude for our potential deflections. Before beginning the above studies in electrostatic coupling, we reproduced some of Mr. Yagi's work with magnetic coupling in order to be sure that the apparatus was being used in the proper way and to accustom ourselves to the necessary manipulations.

Figure 5 shows diagrammatically the arrangement of apparatus for our work, as used with the various connections. We shall call the circuit shunting the arc the primary circuit. Our first experimental problem was to construct an arc which would give us fairly persistent oscillations in our shunt circuit. They had to be steady enough to produce a figure on the Braun tube screen which could be photographed. After many trials of

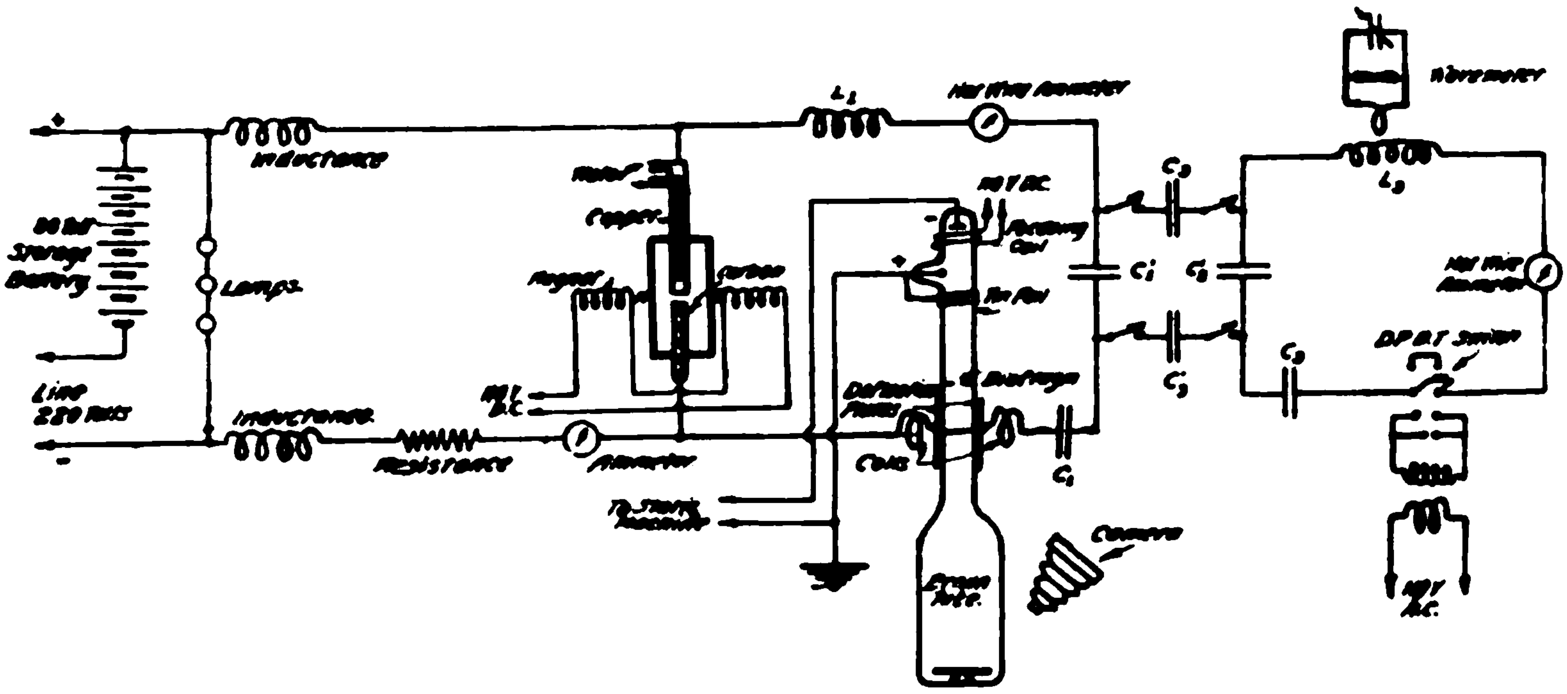

FIGURE 5

different arrangements, and failures, we finally used the enclosed arc shown in the diagram with fair success. The negative electrode was a solid carbon rod about one centimeter (0.4 inch) in diameter with its end filed off flat. This last is necessary to keep the arc from varying in length as it moves around. The carbon should be rotated slowly to prevent the arc from burning away one point on the carbon; but we did not do this, as we found it sufficient to turn the carbon part way around once in a while, or to substitute a new carbon. Sometimes we used a longitudinal magnetic field, which caused the arc to rotate about the axis of the carbon. The transverse magnetic field produced more vigorous oscillations but generally not such steady ones as no field at all.

The arc was enclosed in a porous cup which was properly closed with asbestos, and provided with a peep hole for the adjustment of the arc, and with an alcohol drip. A flame from a bunsen burner kept the porous cup hot so that the alcohol which dripped down on the inside was quickly vaporized. The alcohol vapor seemed necessary, for, as soon as the alcohol gave out the oscillations stopped. This alcohol vapor has the effect of steepening the characteristic curve of the arc.

Our Braun tube did not have some features which we wanted but we used it as it was. There was only one diafram in the tube, tho there should have been two to make the spot on the screen small and clearly defined. A focusing coil placed as indicated in the drawing helped us greatly in obtaining a bright and fairly well defined spot. The strength of the field of this coil and its direction had to be adjusted by experiment for the

best effect. We were troubled by the jumping of the beam, due probably to the accumulation of charges on various parts of the tube, until we partially covered the large end of the tube with tinfoil and put a strip of foil around the tube close to the positive electrode, grounding each of these foils along with the positive electrode. Current deflections were obtained by passing the current thru two coils placed on opposite sides of the tube so that their magnetic fields would add. In our experiments two coils of twenty turns each were used in series. For potential deflections, a couple of metal plates were held as close as possible to the tube and on opposite sides. If our plates had been sealed into the tube we would have been able to get larger deflections for the potentials used—a thing which we often needed. For operating the tube we used a two-plate Wimshurst electrostatic machine driven by a motor.

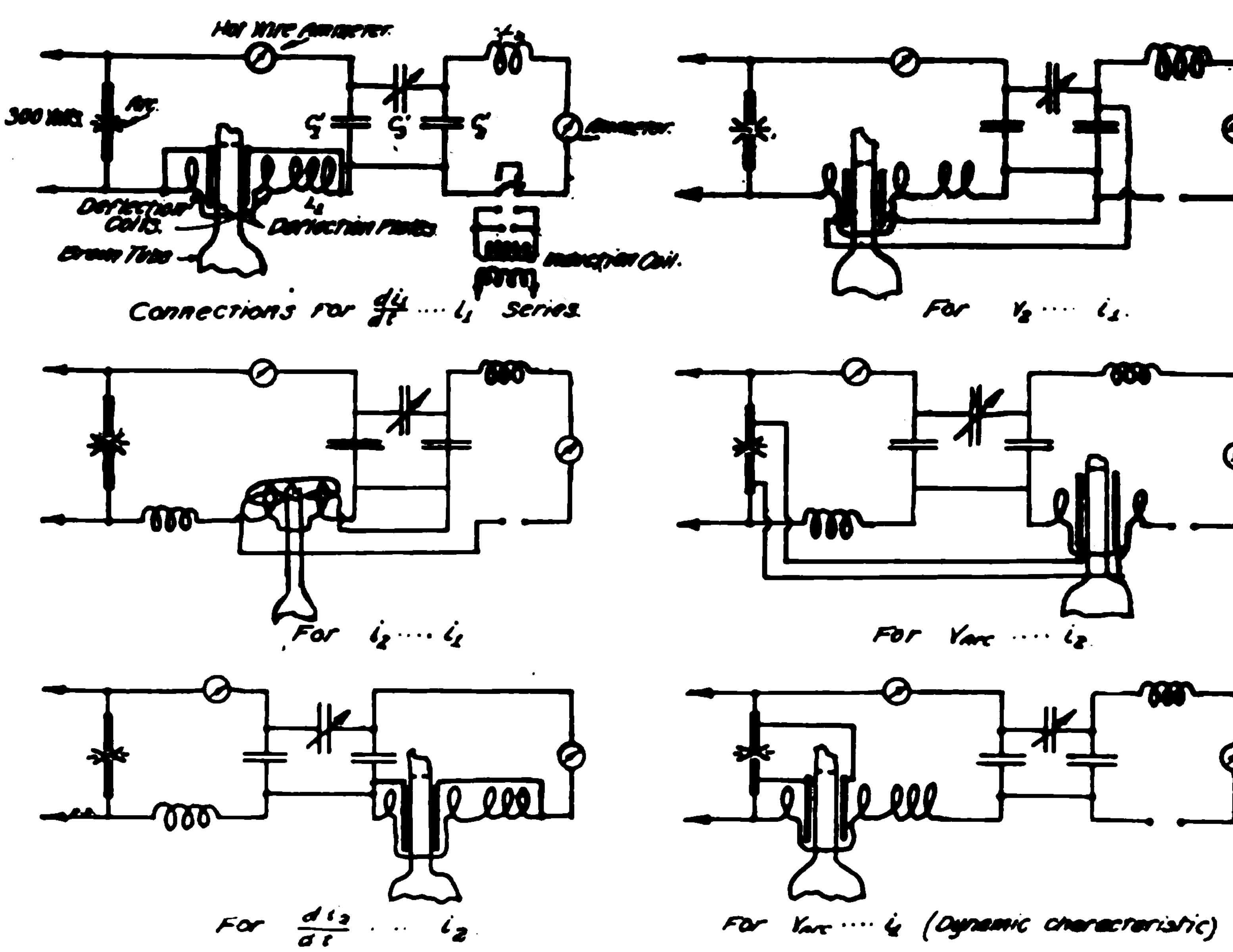

FIGURE 6

Placed a little to one side of the tube, and focused on the front of the screen, we had our camera. With an f.-6.3 anastigmat lens and Cramer's "Isochromatic" plates, an exposure of from one-fifth to three-fifths of a second was sufficient for good

photographs. Once in a while we thought our plates showed the effects of x-rays, probably from the aluminum diafram in the tube.

The ammeters used in the two oscillatory circuits were of the hot wire type, each being calibrated with the line ammeter and direct current. However, as the secondary current meter was burned out just preceding its calibration, it was calibrated, using a wire and shunt as nearly like the original as possible. At any rate, even if the values thereby given are only approximate, we know relative values from the readings.

The inductance in each circuit aside from that in the deflection coils was in the shape of a spiral. The capacity was made up of sections of Murdock molded condensers of approximately 0.0017 microfarads capacity each. The capacities of all sections were assumed equal when the coefficient of coupling was calculated. A variable air condenser, with a capacity at fifty-five scale divisions equal to that of one section of condenser, was used with the coupling capacity to make the coupling continuously variable.

When the switches shown in the secondary circuit are thrown towards the spark gap, we have a means of exciting our secondary for tuning purposes. The wave meter consisted of an inductance in series with a variable air condenser calibrated for wave lengths. A low pressure hydrogen tube was connected across the terminals of the inductance or condenser, to indicate maximum potentials in the circuit.

On the line side of the arc, there was some dead resistance for controlling the current thru the arc, as well as a large inductance in each line to prevent oscillations from the shunt circuit from getting into the line. The inductance in each line was the secondary of a commercial house-supply transformer. As an additional precaution against the oscillations getting back into the line, three incandescent lamps were placed in series across the line.

In our work we found it very convenient to short-circuit C_1, C_3, and C_2, and to use one section of condenser for C_1', and C'_2 each, leaving C_3' for varying the coupling. With this arrangement and with all the capacity we had available, the coupling could be varied from zero to over ninety per cent.

In each case, we were careful to have our circuits tuned so that they satisfied the condition of resonance before coupling. The wave length of an oscillation in the primary circuit, which seemed to be readily reproduced, was determined and the second-

ary was tuned to that, without the coupling, by means of the spark gap and transformer. Now, according to the theory which we have verified in the first part of this work, we should always have this original frequency in each of the circuits no matter what the value of the coupling. Therefore, after coupling, we placed the wave meter near the primary circuit and adjusted the arc until the tube on the meter glowed when the instrument was set for the original wave length.

The same procedure was used in obtaining each series of relations between variables. Having our deflecting coils and plates on the tube properly connected, we began with zero coupling between the circuits and increased to the maximum coupling, taking photographs as we proceeded whenever we got a new figure or a great change in a preceding one. As each photograph was taken, we noted the value of the coefficient of coupling and the primary, secondary, and line currents. For the two current deflections we used two sets of coils at right angles on the tube, which coils, so far as our work was concerned, had practically no mutual inductance (as we found by test).

The following table gives in the rows the series with the same variables while the columns give those figures of the different series with approximately the same coupling. Since a figure generally evolved gradually into the next figure taken in that series we can easily "interpolate" figures to fill out some of our columns, if we care to develop the set of figures with any certain coupling.

CLASSIFICATION OF PHOTOGRAPHS ACCORDING TO COUPLING VALUES

Numbers in table refer to numbers on photographs

Coupling %	0	30	33	42	56	61	65	70	74	78	82	87	90	92
(1) $\dfrac{\frac{di_1}{dt}}{i_1}$	..	..	..	..	13	..	..	..	..	14	15	16	17	18 19
(2) $\dfrac{V_2}{i_1}$	20	21	22	..	23	..	21	24	..	25	26	..	27	28
(3) $\dfrac{i_2}{i_1}$	29	..	..	30	31	..	32	..	33	..	34	35	36	37
(4) $\dfrac{V_{arc}}{i_2}$	38	..	..	39	..	40	41	..	..	..	42 43	..	44	45 46
(5) $\dfrac{\frac{di_2}{dt}}{i_2}$	47	..	..	..	..	48	49 50	..	51	..	52 53	54 55	56	57 58

In Figure 7, we have plotted the current values in the secondary circuit as they varied with the coupling as obtained in series (4). Similar curves were obtained for the other series except in series (1) and (3), where the second maxima seem to be missing.

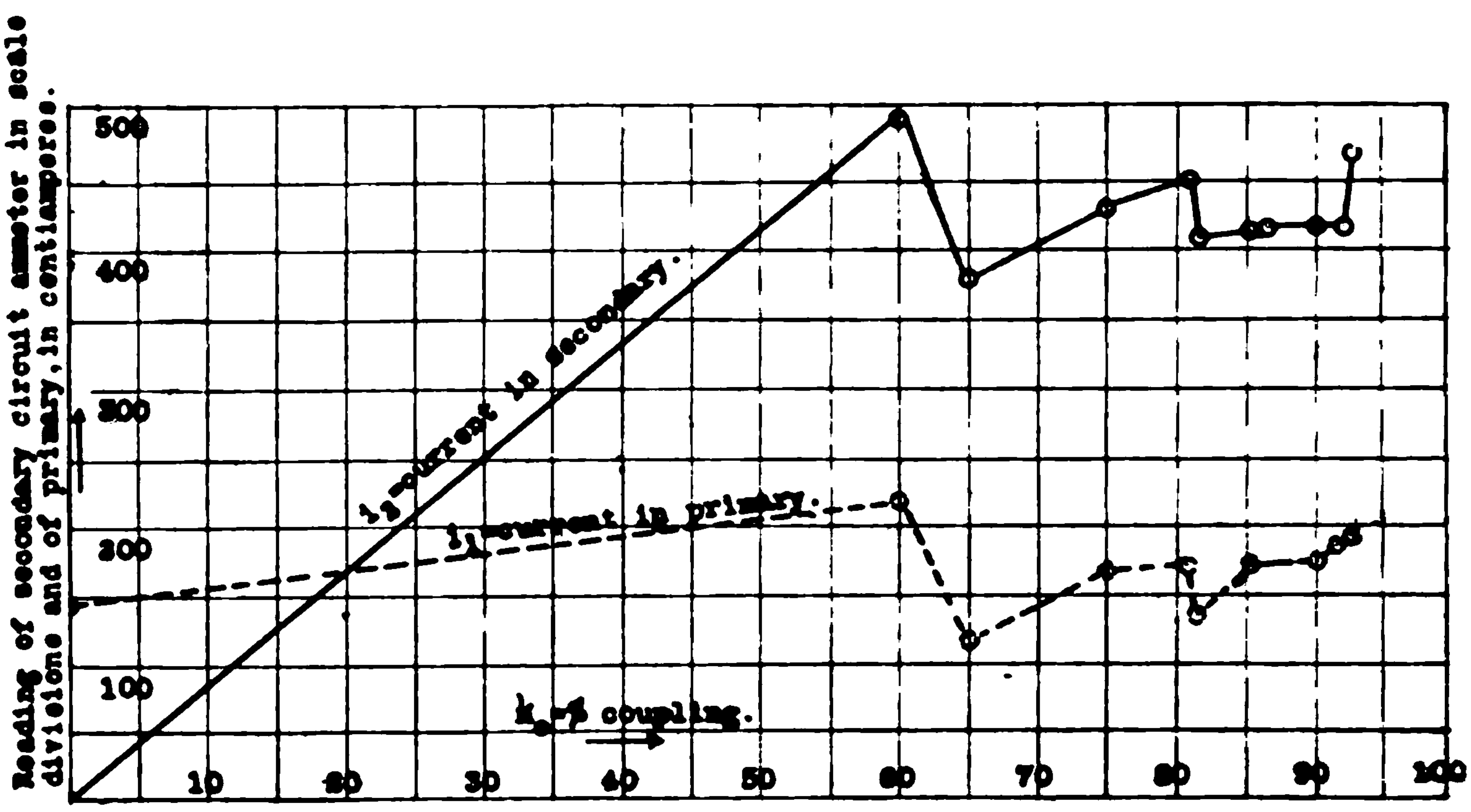

FIGURE 7

Considering Figure 7, the first maximum of current occurs at a place where the value of coefficient of coupling was sixty per cent. Now if the curve (Figure 2), showing $\frac{\lambda''}{\lambda'}$ for various coupling values, be consulted, it will be seen that for $k_e = 60\%$, $\frac{\lambda''}{\lambda'} = 2$. The second maximum on Figure 7 is at $k_e = 80\%$, at which value by curve 1, $\frac{\lambda''}{\lambda'} = 3$. By reference to the photographs of the figures obtained at these values the values of $\frac{\lambda''}{\lambda'}$ obtained above are verified. Therefore, when the ratio of the two frequencies is an integer, the root-mean-square value of the current is a maximum. The variation of the primary current, also shown in Figure 7, leads us to the conclusion as stated for the secondary circuit. Nothing definite can be said about the line current unless it is that it seems to be a minimum when the oscillating current is a maximum.

As series (1) gives us the value of $\frac{di_1}{dt}$ with respect to i_1, it is

useful for the exploration of the variation of currents, and consequently potentials, with respect to time. Let us take figures which were obtained with 82% coupling and develop the curves for time.

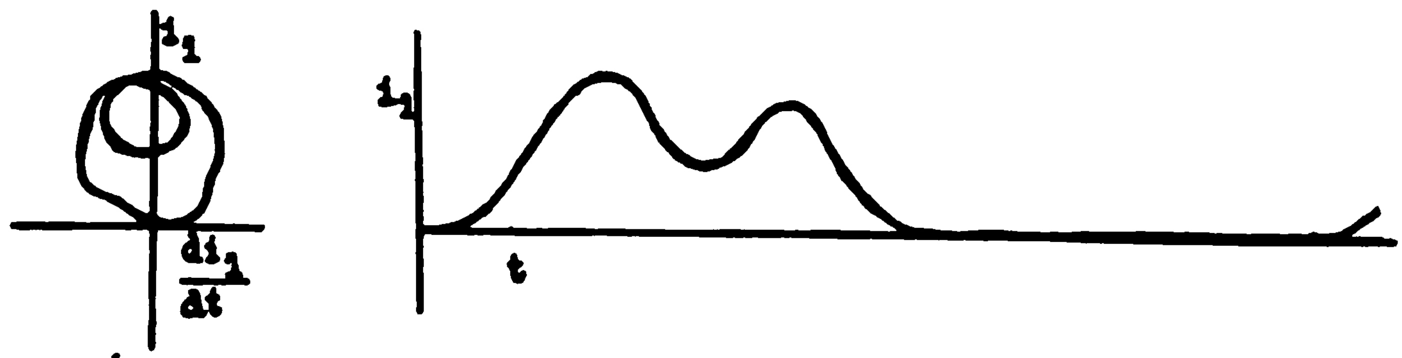

Figure 8.

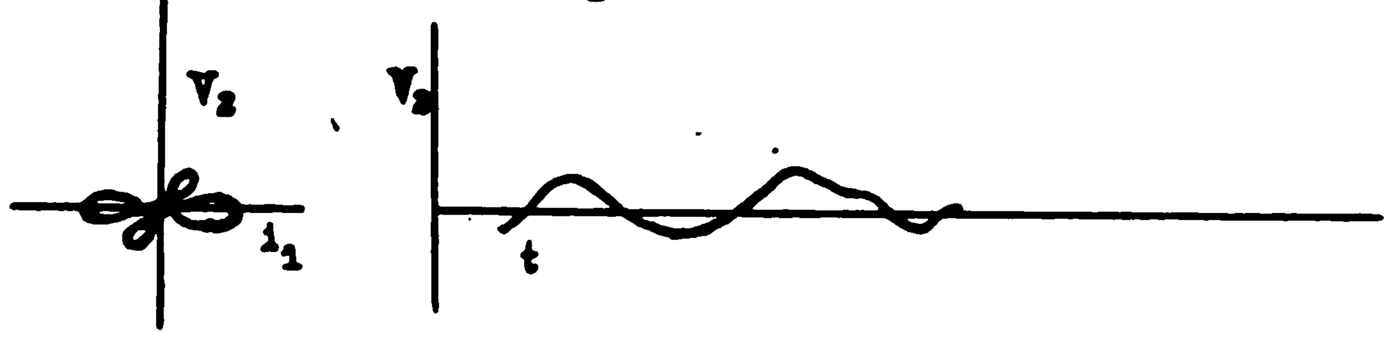

Figure 9.

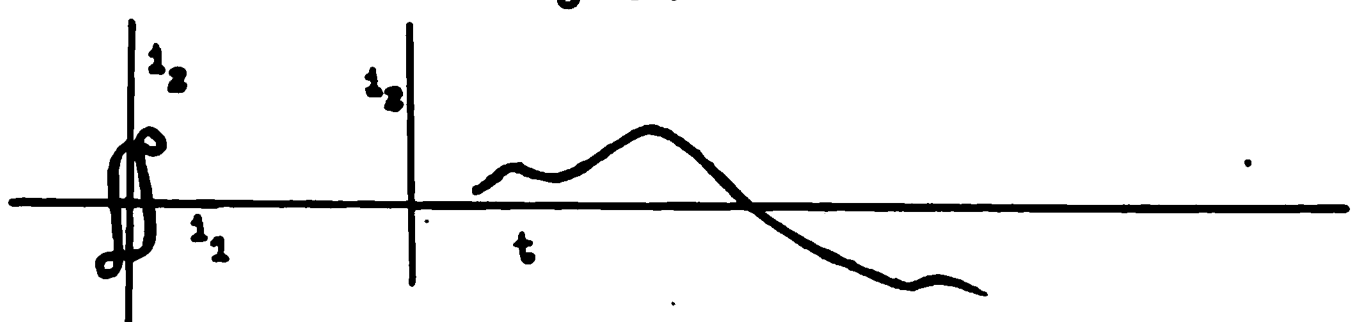

Figure 10.

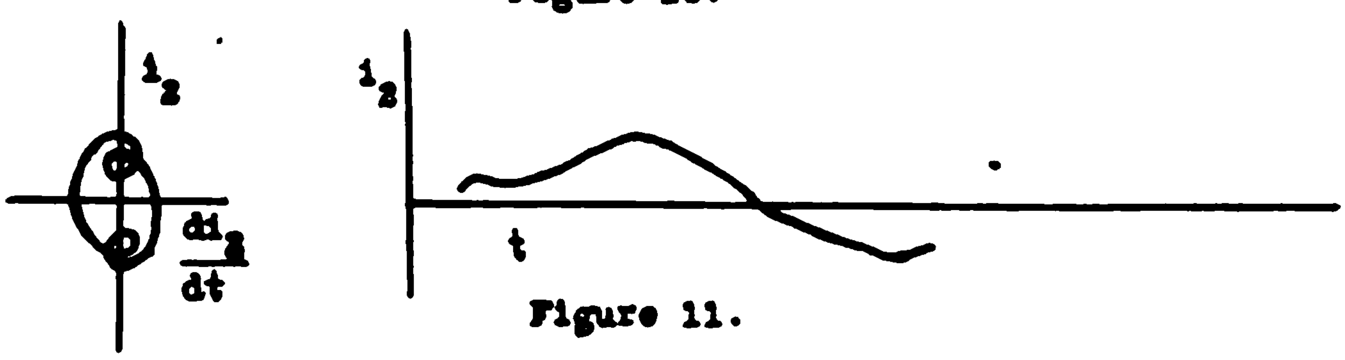

Figure 11.

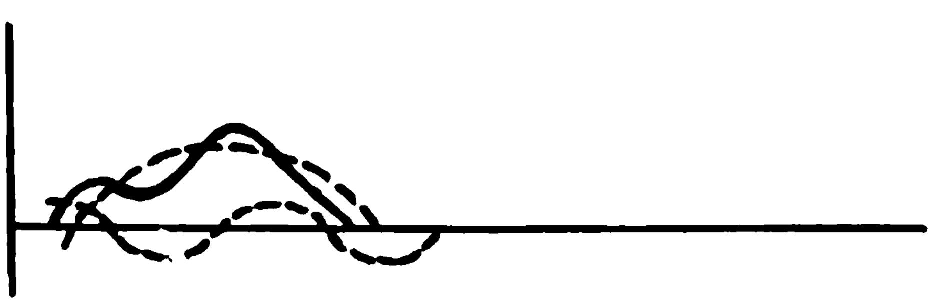

Figure 12.

FIGURES 8 TO 12

Figure 8 shows, first the figure as obtained from the photograph properly placed for development, and then the resulting curve. This is not the actual shape of our wave, as we cannot determine equal intervals of time on our figure, but it does

heretofore.

Figures subsequent to Figure 12 are photographs of the figures produced on the Braun tube screen. The figures of the same series or those with the same coupling value may be picked out by reference to the table on a preceding page.

From this study one is encouraged to believe that electrostatic coupling should have a place in the transference of energy between radio circuits first, because a high degree of coupling is possible and second, because there is practically only one wave in the circuits when such a high coupling value is used.

I desire to thank Mr. Laurens E. Whittemore, of the Physics Department, of the University of Kansas, for his constant and untiring direction and help in this work. To the Department itself, I wish to express my appreciation for the use of apparatus used in this research. This apparatus is shown in Figures 59 and 60.

FIGURES 13 THRU 58

FIGURE 59

Figure 60

SUMMARY: The foregoing paper tells first of an investigation of the mathematical theory of electrostatically coupled circuits, and of the experimental verification of the conclusions drawn; and second, of the study, by means of the Braun tube and sustained oscillations, of the relations existing between the variables in the electrostatically coupled circuits using various values for the coefficient of coupling.

It is found by theory and also by experiment that by one method of tuning, as the circuits are coupled closer, one wave length remains constant while the other approaches infinity, thereby concentrating an increasing proportion of the energy in the one wave. It is shown by curves and by Braun tube figures that when the ratio of the two wave lengths is a whole number, the root-mean-square value of the current is a maximum.

A number of Braun tube photographs are given to show the relations between the various currents and voltages in the circuits.

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PROCEEDINGS

of

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EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

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Correction: On page 159 of the 1918 Proceedings, 7th line from the bottom of the page, change this to read:
"terminals of a current supply by means of a self-inductance"

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THE DESIGN OF POULSEN ARC CONVERTERS FOR RADIO TELEGRAPHY*

By

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PREFACE

Neither time nor space permit a really thoro discussion of the technique of arc design. In this paper, I have endeavored to present those electrical and magnetic features of basic importance and major interest.

Past literature has generally dealt only with the electrical characteristics of the Poulsen arc from the viewpoint of the scientist in the laboratory. I have tried as far as possible to discuss these characteristics from the viewpoint of the designing engineer. This different method of treatment has made it possible to inject new material into the theory of operation.

Those portions of the paper dealing with magnetic matters contain new material also, some of which is of basic importance in the proper and economical proportioning of the bi-polar electro-magnets which have been built in sizes up to 80 tons (72,700 kg.) dead weight.

This paper will be followed by another describing the direct current generating equipment, control apparatus, and other matters of engineering interest connected with high power stations.

At present, a high power station may be defined as one in which the antenna current exceeds 150 amperes under ordinary conditions of ground resistance, and so on. Modern 100 kilowatt converters operate at 150 amperes radiation continuously with a temperature rise not in excess of 40° C. in any part.

HISTORICAL

The negative slope of the volt-ampere characteristic of the direct current electric arc makes it a possible means of obtaining radio frequency currents.

In the early days of radio telegraphy, before the advantages

* Received by the Editor, February 13, 1919.

of continuous waves were generally realized, several types of arcs were devised for producing continuous oscillations, but due to inherent limitations and the difficulties of development, no marked progress was made in their design until 1913, when tests of the United States Navy Department from the Arlington Station showed that continuous waves should be considered very seriously in the radio telegraphy of the future. This created the demand required to expedite development, and the arc operating upon the basic ideas of Valdemar Poulsen has been rapidly developed since that time. No notable advances have been made with arcs of other types. This is probably due to the fact that the mechanical and electrical problems involved are severe, and further because the Poulsen electrical cycle is admirably suited for converting large amounts of direct current electrical energy into radio frequency energy.

The developments of the last five and a half years have advanced this arc from converters of 30 kw. normal full load rating to 1,000 kw. units with 25 per cent. 2-hour overload capacity. Most of this development has occurred within the last 3 years.

General

The theory of the operation of the Poulsen cycle has been studied by a large number of investigators. Very complete bibliographies of this literature are given in Zenneck's "Lehrbuch der Drahtlosen Telegraphie" and Pedersen's article "On the Poulsen Arc and Its Theory," Proceedings of the Institute of Radio Engineers, volume 5, number 4.

The theory of operation described herein is based upon the theories of many investigators, notably those of Barkhausen and Pedersen, combined with certain conclusions of the writer.

The Poulsen Arc Converter Cycle

The arc radio frequency cycle may be divided into two halves. The first is that during which the radio frequency current I_s is circulated by emf. set up by energy stored in the L and C. The second is the energy adding period during which I_s is circulated by emf. set up by energy from the d.c. supply circuit. This may be termed the "charging period."

Referring to Figure 1, the starting of the d.c. generator charges the condenser C_s to the potential E_d. When the arc is struck, direct current I_d flows thru it in the direction of the arrow, forming one component of the arc current I_a. The

other component is the discharge current I_s from the condenser C_s, thus:

$$I_d = i_a - i_s \text{ (instantaneous values)} \tag{1}$$

Due to the fact that arc flame conductivity is dependent upon gas ionization, it is dependent upon gas temperature, and hence I_a, the current thru the arc. Thus as i_s and i_a increase

FIGURE 1 —Diagram of Circuits

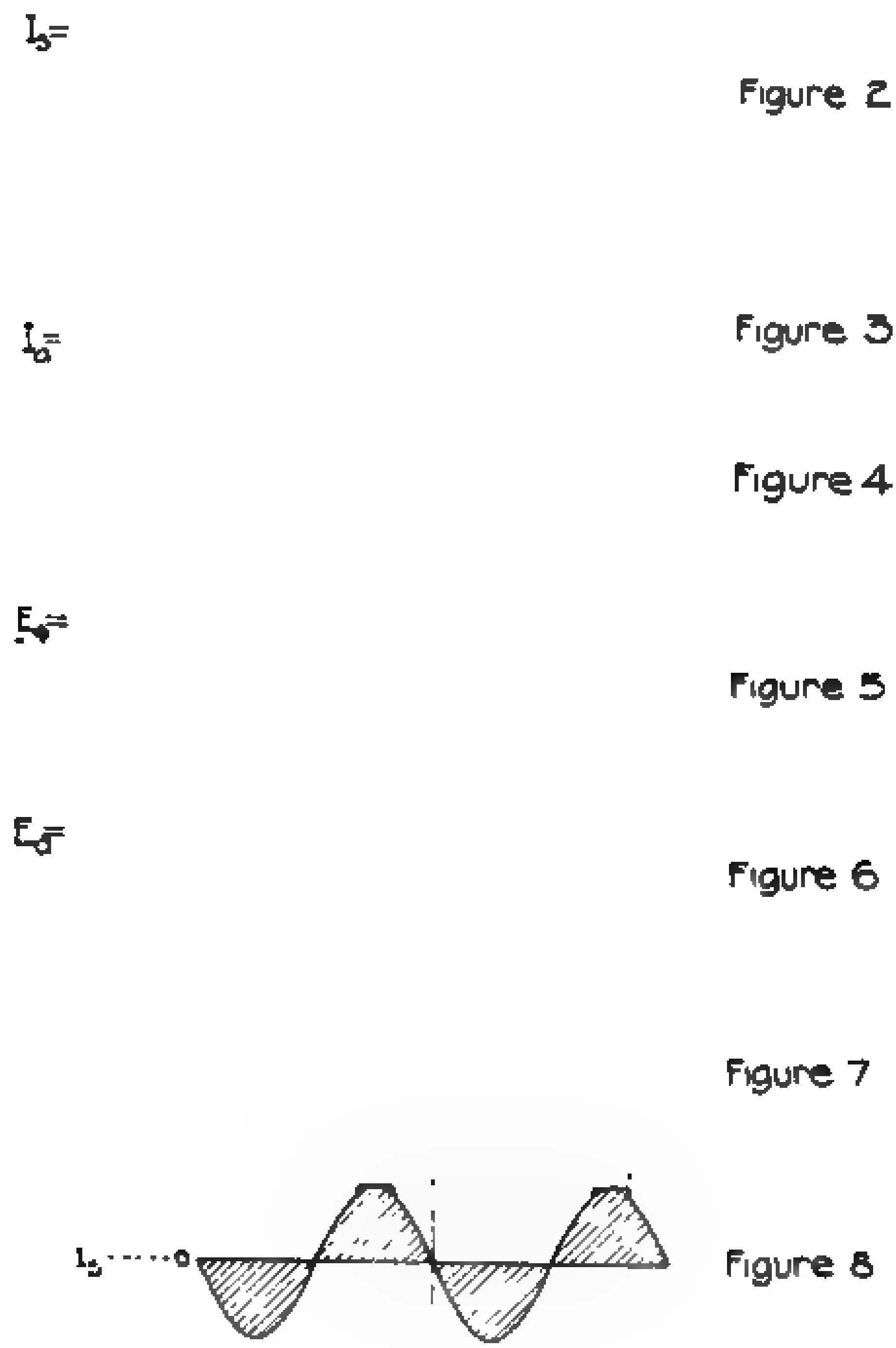
as shown in Figures 2 and 3, the conductivity of the arc flame is raised. This causes a further increase in i_s until finally the peak of the i_s curve is reached at "*b*." At this point the energy which, at the beginning of the cycle was stored in C_s as potential energy, has been completely changed to kinetic energy stored in the magnetic field of the inductor L_s. The condenser charge is therefore zero, and the currents i_s and i_a are a maximum.

The magnetic field of L_s now begins to collapse. This continues to make current flow in the same direction. The process continues until the point "*c*," Figure 2, is reached. Condenser C_s is now fully charged in the polarity opposite to its initial charge, and the energy in the oscillatory circuit is once again in the potential form.

The condenser now begins to discharge, and the second half of the radio frequency cycle begins. During this half cycle, energy from the d.c. circuit is supplied to the oscillatory circuit. Current leaving the condenser does not pass up thru the arc,

forming a portion of I_a as in the preceding half cycle, but passes thru the d.c. circuit in accordance with the equation:

$$I_d = i_a + i_s \text{ (instantaneous values)} \tag{2}$$



FIGURES 2–8

As i_s increases toward "d," Figure 3 shows that i_a approaches zero, and at "m" it has been reduced to such a low value that the stream of ions forming the arc flame starts to rupture under the influence of the magnetic field. This continues to the point "o," at which the arc is completely extinguished and i_a is zero. Thus:

$$I_d = i_s \text{ (instantaneous values)} \tag{3}$$

The next instant i_s decreases from "c" toward the point "n." Therefore there is a slight reduction in I_d (Figure 7) which induces an emf. E_{Ld} between the terminals of the inductance L_d in the d.c. circuit. This surge has a much steeper wave front than the sinusoidal radio frequency oscillations and is unable to force its way beyond the first few turns of L_s. The resultant increase in voltage across the arc is sufficient to jump the gap between the electrodes and re-establish i_a. This occurs at "n," and more and more I_d is shunted off thru the i_a path as i_s approaches zero at "e."

The point "e" is at the beginning of a second cycle identical with that just described, with the exception that, whereas at "a," the potential E_c across C_s was only that of E_d, at "e" it has been augmented by the discharge of L_s also. Thus, when the arc is first started, there is a transient period extending over several cycles, during which the peak of E_c for each succeeding cycle is constantly increased until a stable condition is reached, which depends solely upon the resistance of the radio frequency circuit, all other conditions remaining constant. Thereafter the effective value of E_c may be computed by the well known equation:

$$E_c = \frac{I_s}{2\pi f C_s} \tag{4}$$

THE ARC VOLTAGE E_a

Altho I_s is sinusoidal and I_a is a sinusoidally pulsating unidirectional current, the voltage across the arc, E_a, has a jagged wave form. When i_s is at "a" and the arc is struck by bringing the electrodes together, e_a takes a certain value as shown. Due to the drop in arc flame resistance produced by increasing current and because the flame resistance drops at a rate greater than the first power of the current, e_a, which equals $r_a\, i_a$, decreases with an increase in i_a as previously described. This is the reason for the dip in the e_a curve and illustrates the well known falling characteristic of the arc. As i_a approaches zero, e_a increases up to the extinction point "m." Then comes re-ignition at "n" and e_a drops as gap ionization increases. The cycle then repeats itself.

As Pedersen points out, there is not necessarily much difference in the amplitude of the voltage peaks "m" and "n" because the arc is burning from points back on the electrodes at "m" and is, therefore, long, while at "n" the voltage is only that

necessary to jump the gap between the electrodes at their nearest point.

The ignition voltage at "n" is of course dependent upon the ionization in the gap at that time. This ionization is controlled by the magnetic field strength, but inasmuch as the field has had the opportunity of scavenging the gap for a time prior to ignition, slight changes in its strength are not likely to make as much difference in the amplitude of the voltage peak at "n" as at "m," because during the period leading up to the peak "m" the gap has been constantly supplied with new ions which were blown out of it by the magnetic field. Hence *slight* changes in field strength probably make a greater difference in the extinction voltage than in that of ignition.

During the period "$a\ b\ c$," Figure 2, e_a at any instant equals $r_a\ i_a$ where r_a is the varying resistance of the arc. No simple law is followed from "c" to "e," because of the points "m" and "n." During the period "$a\ b\ c$," e_a opposes i_s, which is a component of i_a. For maximum I_s it is, therefore, desirable to have E_a a minimum during this period. However, thru "$c\ d\ e$" it is desirable to have the effective value of the E_a wave a maximum, provided the "m" and "n" peaks do not cause too great distortion. Thus for a complete cycle the effective value of the E_a wave useful in circulating I_s is the effective value during the period "$c\ d\ e$" minus the effective value during the period "$a\ b\ c$."

The equivalent sine wave of in-phase emf., which we may term E_ϕ, is that which has the same effective value as the difference between the effective values of the two halves of the E_a curve above mentioned. This is obviously the voltage drop across R_s, that is $R_s\ I_s$. It is shown in Figure 5.

The peak value of $e_\phi = E_{da}$ ($E_{da} = E_d$ minus the $R\ I$ drop in the d.c. circuit) because experiments show that when the magnetic field strength is of the proper value and the arc is operating under good conditions its effective direct current resistance, $\dfrac{E_{da}}{I_d}$ equals the resistance R_s of the oscillatory circuit.

Since $$E_\phi = R_s\ I_s,$$

substituting $$E_\phi = \frac{E_{da}}{I_d} \cdot \frac{I_d}{\sqrt{2}} = \frac{E_{da}}{\sqrt{2}}$$

Hence $$E_{da} = e_\phi \text{ (peak)} \qquad (5)$$

Figures 9 and 10 give experimental proof of the foregoing. They are plotted from data taken at two high power stations.

Altho the wave form of E_a is not sinusoidal, it is possible to

resolve it into components which may be easily treated analytically.

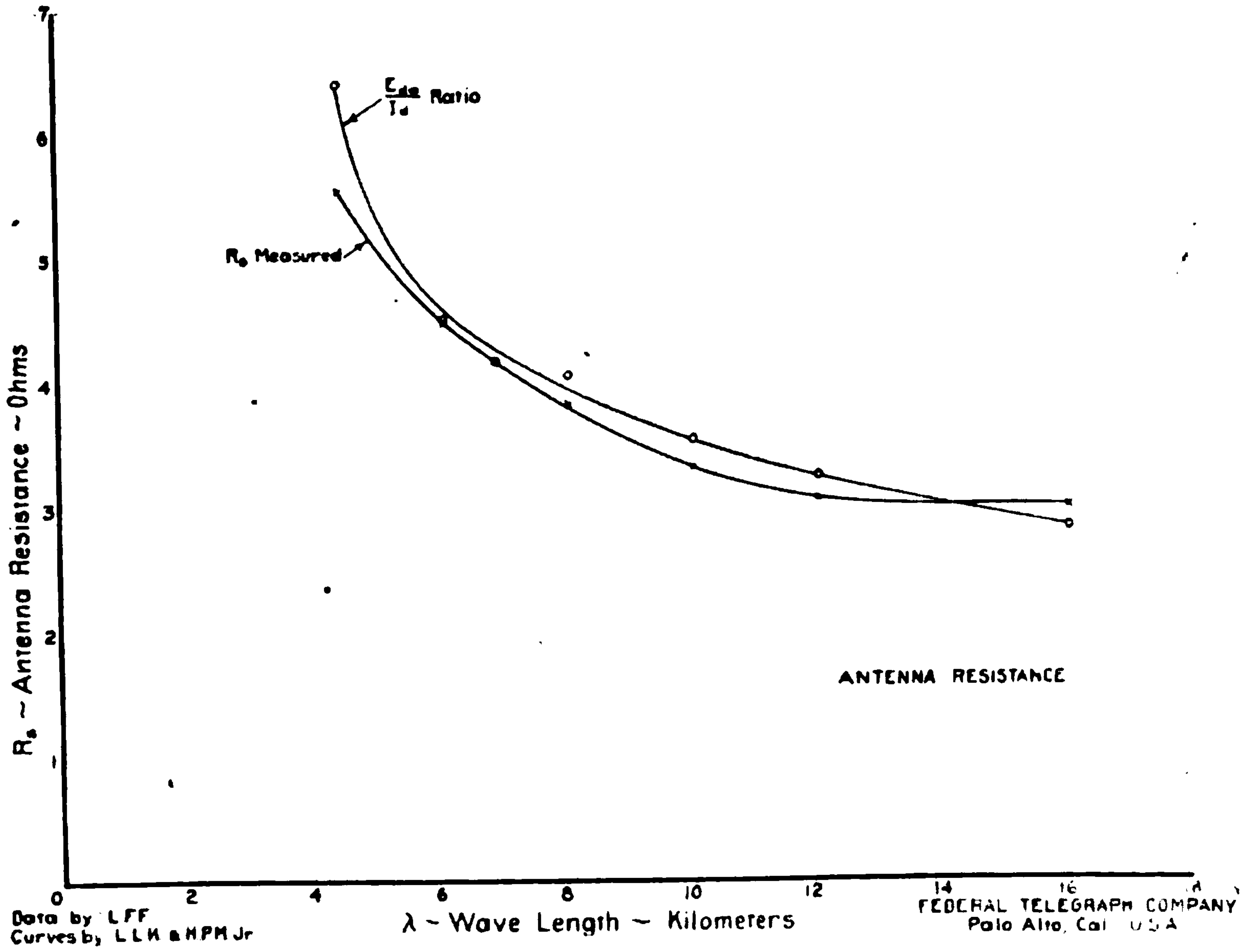


FIGURE 9

These components are:

(1) E_{da} maintained by the d.c. generator.

(2) A sinusoidal uni-directional pulse occurring once per radio frequency cycle.

Figure 6 shows the superposition of these components. As previously stated, the difference between the effective values of the half cycles of this wave must be 0.707 E_{da}. When this is the case the effective value of

$$E_a = 1.4\, E_{da} \tag{6}$$

This is shown in the mathematical analysis which follows and is also proven experimentally.

Referring to Figure 11:

Let e_a = instantaneous voltage across arc.

$$E_a = \text{effective value of } e_a = \left[\frac{1}{2\pi}\int_o^{2\pi} e_a^2\, d\,\theta\right]^{\frac{1}{2}}$$

E = maximum value of uni-directional pulse that occurs once every cycle of the radio frequency current.

e = instantaneous value of pulse $= E \sin \theta$

E_{da} = d.c. component of arc voltage.

$$e_a = E_{da} + e = E_{da} + E \sin \theta$$

$$e_a^2 = E_{da}^2 + E^2 \sin^2 \theta + 2 E_{da} E \sin \theta$$

$$\frac{1}{2\pi}\int_0^{2\pi} e_a^2 \, d\theta = \frac{1}{2\pi}\int_0^{2\pi} E_{da}^2 \, d\theta + \frac{E^2}{2\pi}\int_0^{\pi} \sin^2 \theta \, d\theta + \frac{2 E_{da} E}{2\pi}\int_0^{\pi} \sin \theta \, d\theta$$

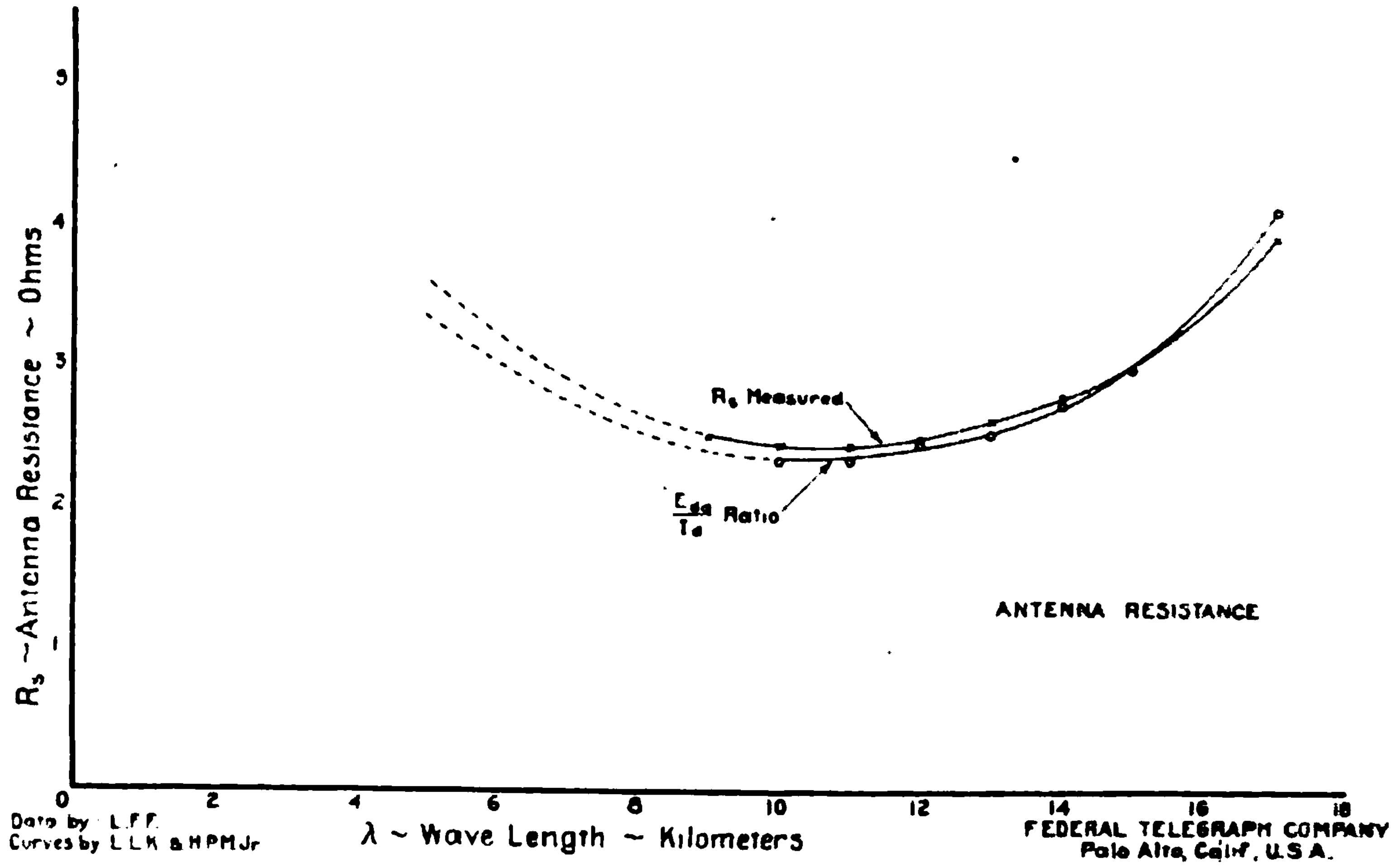


FIGURE 10

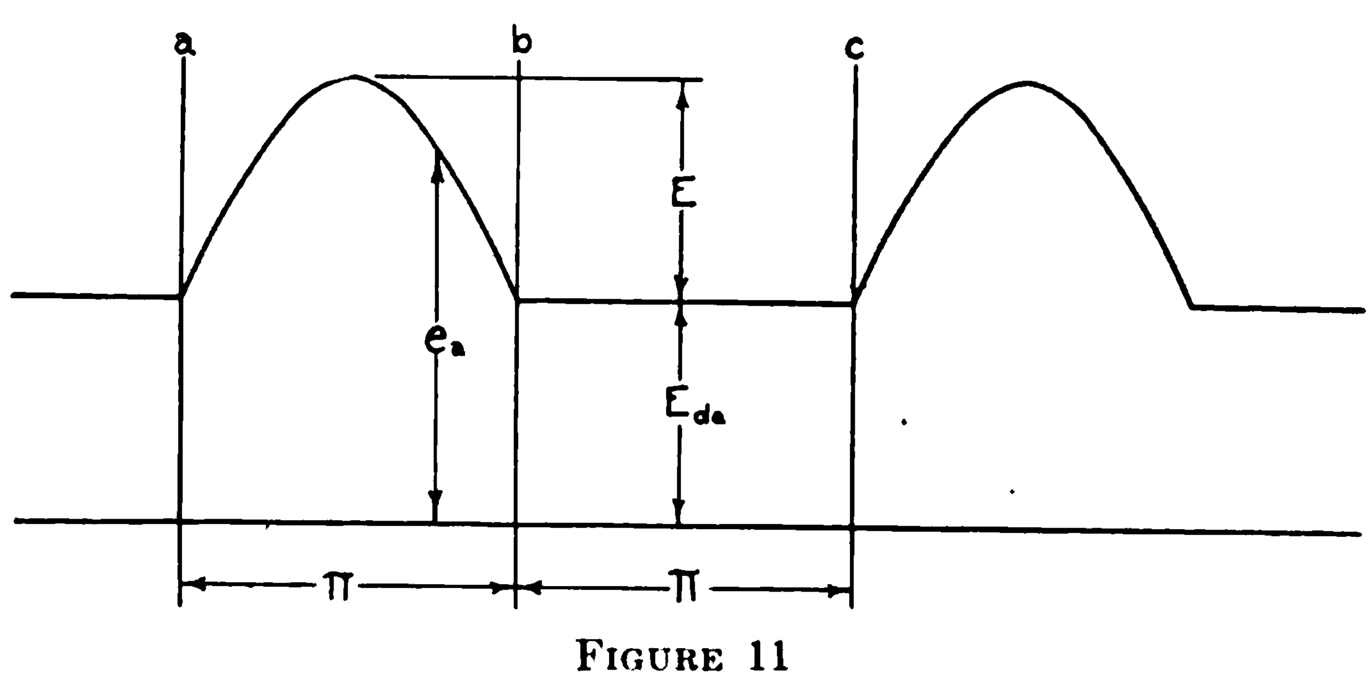


FIGURE 11

Integration of Parts

$$\frac{1}{2\pi}\int_o^{2\pi} E_{da}^2\, d\theta = \frac{E_{da}^2}{2\pi}\int_o^{2\pi} d\theta = E_{da}^2$$

$$\frac{E^2}{2\pi}\int_o^{\pi} sin^2\theta\, d\theta = \frac{E^2}{4\pi}\int_o^{\pi}(1-cos\,2\,\theta)\, d\theta = \frac{E^2}{4}$$

$$\frac{E_{da}E}{\pi}\int_o^{\pi} sin\,\theta\, d\theta = \frac{2\,E_{da}E}{\pi}$$

$$\therefore \frac{1}{2\pi}\int_o^{2\pi} e_a^2\, d\theta = E_{da}^2 + \frac{E^2}{4} + \frac{2}{\pi}E_{da}E$$

$$E_a = \sqrt{E_{da}^2 + \frac{E^2}{4} + \frac{2}{\pi}E_{da}E}$$

The value of E_a during the half cycle "$b\ c$," Figure 11, is E_{da}. During the preceding half cycle "$a\ b$" the value of E_a may be derived as follows:

$$e_a^2 = E_{da}^2 + E^2 sin^2\theta + 2\,E_{da}E\, sin\,\theta$$

$$\frac{1}{\pi}\int_o^{\pi} e_a^2\, d\theta = \frac{E_{da}^2}{\pi}\int_o^{\pi} d\theta + \frac{E^2}{\pi}\int_o^{\pi} sin^2\theta\, d\theta$$

$$+\frac{2\,E_{da}E}{\pi}\int_o^{\pi} sin\,\theta\, d\theta = E_{da}^2 + \frac{E^2}{2} + \frac{4}{\pi}E_{da}E$$

$$E_a \text{ (half cycle "}a\,b\text{")} = \sqrt{E_{da}^2 + \frac{E^2}{2} + \frac{4}{\pi}E_{da}E}.$$

For the difference between the two half cycles "$a\ b$" and "$b\ c$" of e_a to equal $0.707\,E_{da}$,

$$\sqrt{E_{da}^2 + \frac{E^2}{2} + \frac{4}{\pi}E_{da}E} - E_{da} = 0.707\,E_{da}$$

$$E_{da}^2 + \frac{E^2}{2} + \frac{4}{\pi}E_{da}E = 2.91\,E_{da}^2$$

$$E^2 + \frac{8}{\pi}E_{da}E - 3.83\,E_{da}^2 = 0$$

$$E = \frac{-\frac{8}{\pi}E_{da} \pm \sqrt{\frac{64}{\pi^2}E_{da}^2 + 15.31\,E_{da}^2}}{2}$$

$$E = E_{da}\cdot\frac{-2.54 \pm 4.67}{2} = 1.07\,E_{da}$$

The $(-)$ minus value of 4.67 in the preceding equation is disregarded. It evidently gives the value of $-E$ necessary to satisfy the conditions, but which is of no interest in this analysis.

For the particular case when the difference between successive half cycles of e_a is 0.707 E_{da}, the value of E_a in terms of E_{da} is now obtained by substitution from the two equations.

$$E = 1.07\, E_{da}$$

$$E_a = \sqrt{E_{da}^2 + \frac{E^2}{4} + \frac{2}{\pi} E_{da} E}$$

$$E_a = \sqrt{E_{da}^2 + \frac{(1.07)^2 E_{da}^2}{4} + \frac{2.14\, E_{da}^2}{\pi}}$$

$$E_a = E_{da} \sqrt{1 + 0.284 + 0.679} = E_{da} \sqrt{1.963}$$

$$E_a = 1.40\, E_{da}$$

Experimental proof of equation 6 is given by the data taken at Palo Alto plotted on Figure 12.

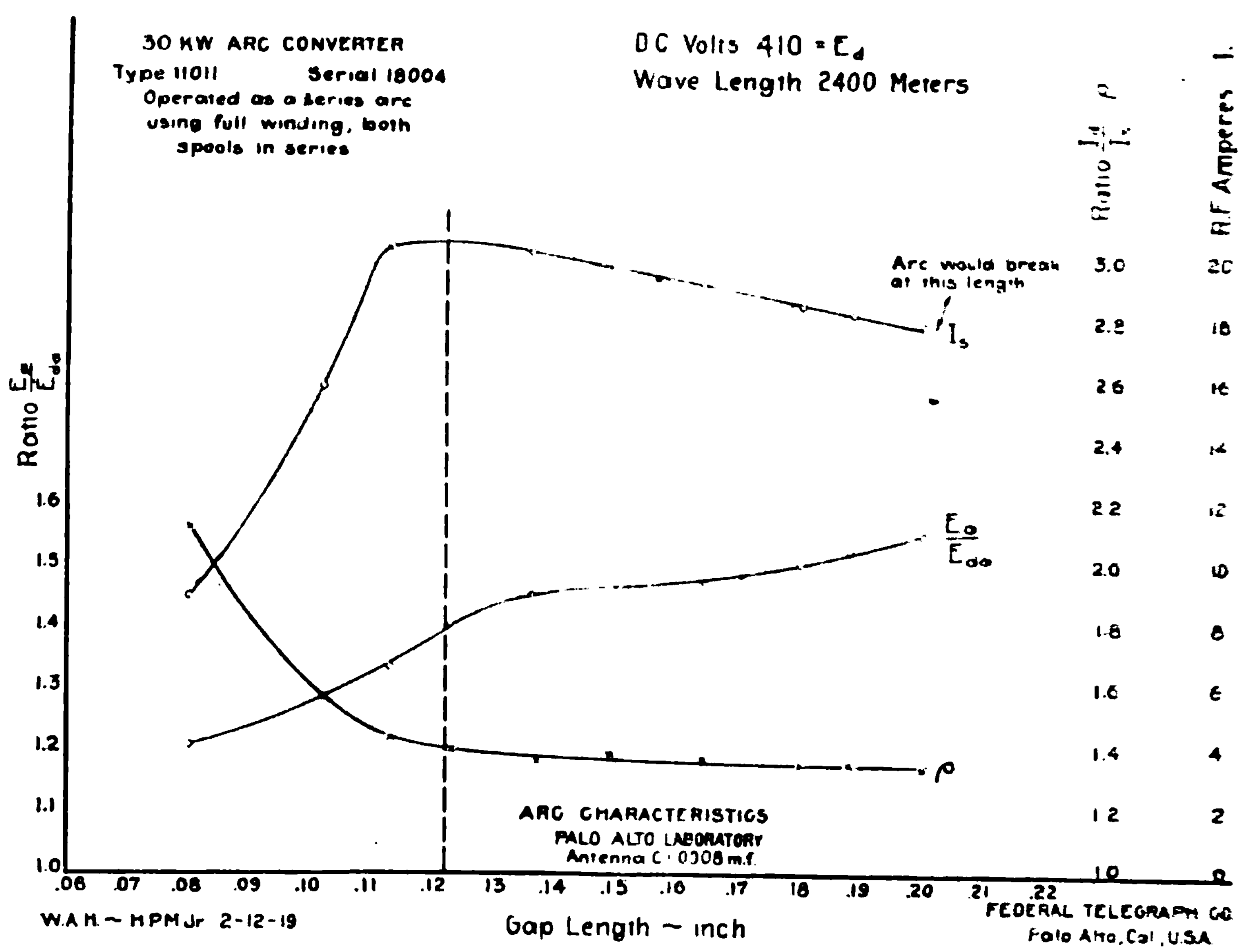


FIGURE 12

It is apparent that when $\rho = \sqrt{2}$, the ratio $\dfrac{E_a}{E_{da}} = 1.4$.

The values of E_{da}, I_d, and I_s were determined with the usual instruments in the usual manner.

E_a was measured by connecting a radio frequency voltmeter,

consisting of a hot wire milliammeter in series with a non-inductive resistance across the radio frequency terminals of the arc. This voltmeter was calibrated on d.c.

The pulsating component of the E_a uni-directional equivalent wave, Figure 6, contains the two peaks "*m*" and "*n*."

It was stated under the heading of "The Poulsen Arc Converter Cycle" that the peak "*n*" was caused by the inductive discharge of L_d and that this pulse of emf. had a steeper wave front than the radio frequency oscillations. These facts have been proven by the following experiments:

Since the voltage E_{Ld} is produced by slight pulsations in I_d in the manner previously described, these cause the collapse of a portion of the air leakage field about L_d. This occurs once every radio frequency cycle, and it is possible to detect the flux changes by placing a wave meter exploring coil in the air in the vicinity of L_d and tuning the wave meter to resonance with the radio frequency. In performing this experiment great care must be taken to make sure that the wave meter ammeter deflection is due solely to the E_{Ld} flux changes and is not due to direct induction from any nearby conductors carrying radio frequency currents or to small radio frequency currents leaking back thru L_d to the d.c. generator.

The extremely steep wave front of E_{Ld} is proven by sphere gap measurements of the voltage between turns of L_s. It is found that the voltage between the end turns next to the arc is higher than between any other turns in the coil. If L_s is not sufficiently large this pulse may carry thru into the condenser C_s. In this case harmonics are set up in the oscillatory circuit.

Figure 8 shows the resultant distortion of the I_s wave. These harmonics will not occur in the antenna current if L_s is sufficiently large. Thus, in practice, a station with a high capacity antenna operating upon short wave lengths is more inclined to have harmonics than would be the case were the L/C ratio higher. As a rule these disturbances are entirely choked back by the end turns of L_s next to the arc.

The Effect of B_g Upon Extinction and Ignition

The theory of the effects of B_g upon extinction and ignition as outlined below is based upon the effect of changes in B_g upon I_s. It is to be understood that no experimental means of localizing and specifically measuring the amplitudes and phase relations of the extinction and ignition voltages has been used.

The magnetic field strength B_g controls both the amplitude and timing of the extinction and ignition voltages.

For any given set of conditions there is a value of B_g which gives optimum I_s. When the arc fields are adjusted to this value, they are said to be "tuned," and the flux density is denoted by β_g.

When B_g is less than β_g, the rate at which ions are removed from the gap is below normal, and hence gap ionization is above normal. This decreases the effective value E_a, which may be proven experimentally by use of the radio frequency voltmeter previously described. Such a condition reduces both the extinction and ignition voltages, and because of the high peak values of these, reduces the effective value of E_a, Figure 4, thruout the "*c d e*" period to a greater extent than thruout the "*a b c*" period. Hence the effective value of the in-phase driving voltage, E_ϕ, Figure 5, is reduced.

The fact that this is the case is easily demonstrated experimentally by lowering the field strength of an arc while it is in operation. The current I_s is immediately lowered.

Conditions with B_g greater than β_g are not altogether the converse of those with B_g less than β_g. This is because, altho the extinction and ignition voltages are abnormally high when B_g is greater than β_g, the time of extinction is advanced and ignition is delayed. This tends to separate the points "*m*" and "*n*" and to foster harmonics. Such improper timing of "*m*" and "*n*" causes a reduction in the amount of energy transferred to the oscillatory circuit and a corresponding reduction in I_s, because the e_a wave form becomes so radically different from the e_ϕ wave of in-phase voltage.

Summing up the foregoing, it is seen that with field strength β_g, the points "*m*" and "*n*," Figure 4, have certain amplitudes and time phase relations with respect to i_s. If B_g is less than β_g, the effective values of both E_a and E_ϕ are reduced and I_s is correspondingly reduced. On the other hand if B_g is greater than β_g, the harmonics in the voltage circulating I_s are augmented and this reduces the energy transferred to the radio frequency oscillations of fundamental frequency. The current I_s is accordingly again reduced.

The Arc Current I_a

The arc current is made up of the two components I_d and I_s. Figure 13 shows the vector relations of the currents involved. I_s is laid off as a base vector of unit length. I_d is laid off at right

angles because it may be considered a sine wave of different frequency, that is, extremely low frequency. Its length is $\sqrt{2}$, since from Figures 2 and 3 it is evident that the crest of the I_s wave equals I_d, and with sinusoidal wave form the ratio of peak to effective values is $\sqrt{2}$.

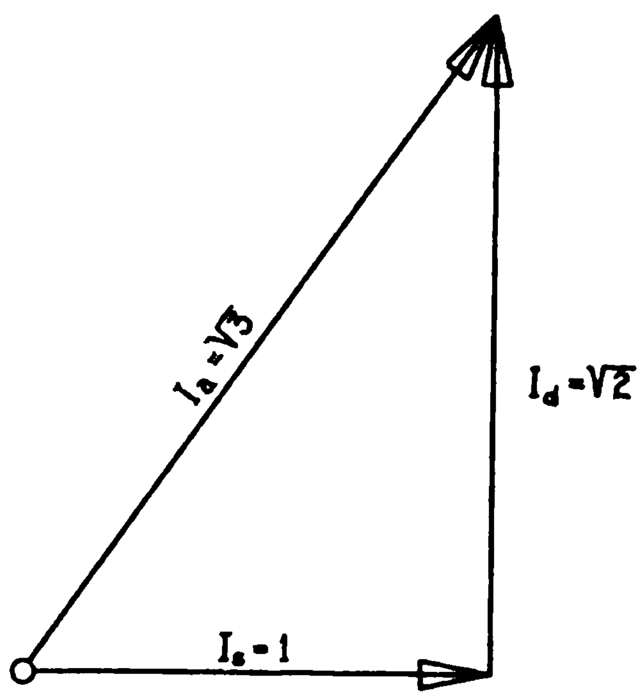


FIGURE 13

The triangle is then closed by $I_a = \sqrt{3}$.

Experimental proof of these current relations is easily obtained by inserting ammeters in an arc circuit to measure I_d, I_a, and I_s (see Figure 1). When the magnetic field strength is of proper value, it is found that

$$I_d = \sqrt{2}\, I_s \tag{7}$$

and

$$I_a = \sqrt{3}\, I_s \tag{8}$$

THE POULSEN CYCLE EFFICIENCY

The efficiency of the Poulsen cycle may be computed from the following:

$$\text{Arc output} = R_s I_s^2 = \frac{E_{da}}{I_d} \times \frac{I_d^2}{\rho^2} = \frac{E_{da} I_d}{\rho^2},$$

$$\text{where } \rho = \frac{I_d}{I_s}$$

$$\text{Arc input} = E_{da} I_d$$

$$\therefore \text{ Arc efficiency } \varepsilon = \frac{1}{\rho^2} \tag{9}$$

$$\text{If } \rho = \sqrt{2}, \text{ then } \epsilon = 50\%.$$

This is the maximum Poulsen cycle efficiency, and the high-

est theoretically obtainable. It corresponds to the Carnot cycle in thermodynamics. If the magnetic field strength is too weak, ρ will be greater than $\sqrt{2}$ and $\frac{E_{da}}{I_d}$ may be greater than R_s. Under these conditions true arc efficiency cannot be determined unless R_s is actually known. However, when the arc magnetic field is tuned, $\frac{1}{\rho^2}$ is a very fair approximation to true arc flame ϵ.

This is shown by Figure 14, which is plotted from data taken at the San Diego (California) High Power Naval Radio Station.

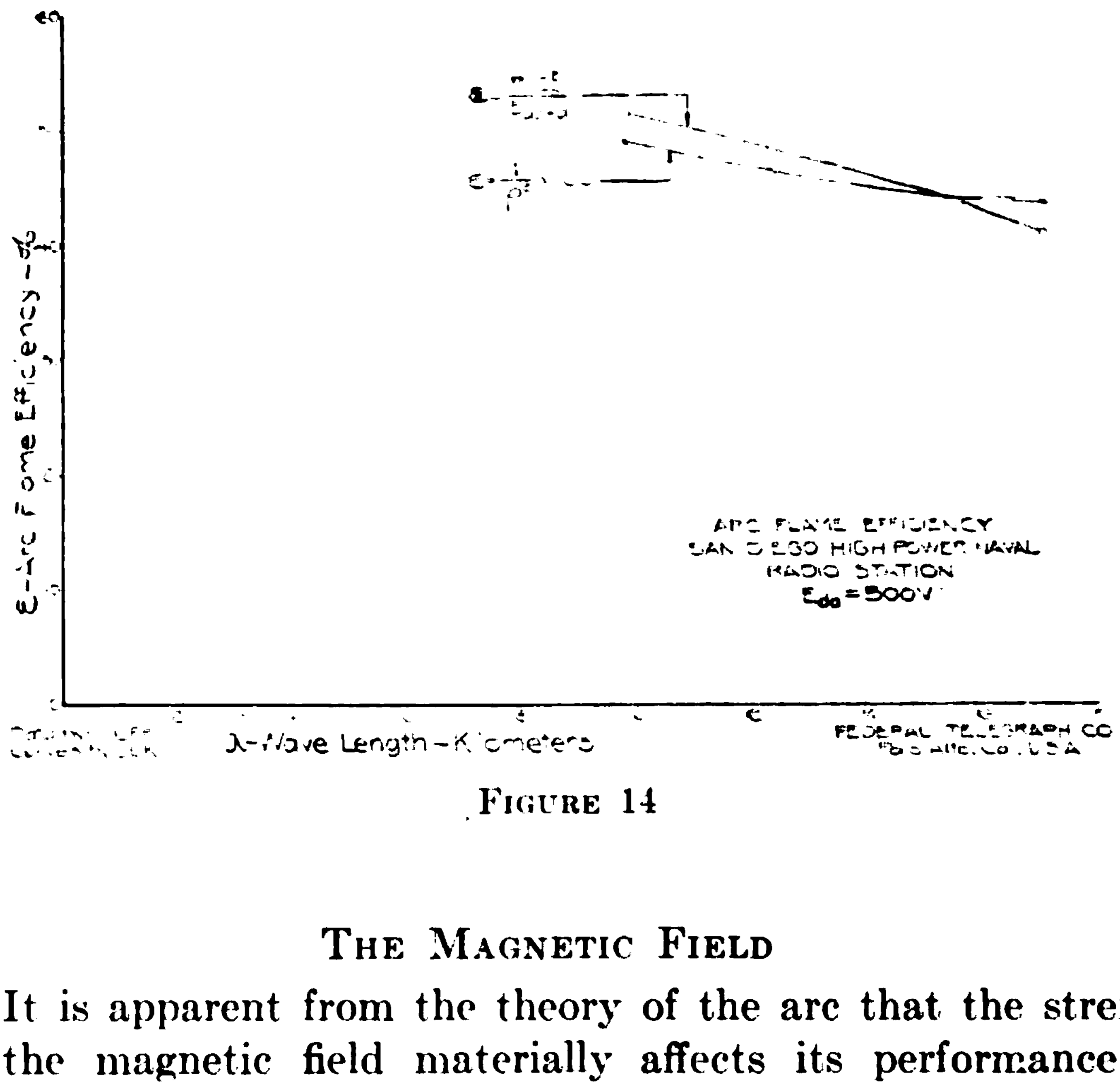


FIGURE 14

THE MAGNETIC FIELD

It is apparent from the theory of the arc that the strength of the magnetic field materially affects its performance. A fraction of each radio frequency cycle is allowed for the extinction of the arc and the scavenging of the gap between the electrodes. The length of this time is inversely proportional to frequency. Therefore, all other factors remaining constant, the strength of the magnetic field required for properly de-ionizing the gap is directly proportional to frequency or inversely proportional to wave length. This is because the rate at which the ions are moved is dependent upon field strength.

Therefore: $$\beta_g \propto \frac{1}{\lambda} \qquad (10)$$

where "λ" is the wave length.

The molecular velocity of the atmosphere in which the arc burns controls the value of β_g. To extinguish the arc properly and de-ionize the gap in the time available, there is no gain in raising the ions above the velocity necessary to break up the ionic stream in the time allowed. Hence, if the molecular velocity of the gaseous medium surrounding the arc is high, it is unnecessary for the magnetic field to increase the velocity of the ions as much as would be the case were their velocity lower. Therefore, the necessary field strength is inversely proportional to the molecular velocity of the gaseous medium surrounding the arc. That is:

$$\beta_g \propto \frac{1}{v} \tag{11}$$

where "v" is the molecular velocity of the gas.

The temperature of the arc flame is so high compared with the temperature of the gases in the chamber when the arc is not in operation that no appreciable error is introduced by the assumption that the absolute temperature of the arc flame is proportional to the power input $E_{da} I_d$. Since the velocity of the molecules of a given gas is proportional to the square root of the absolute temperature of the gas, it follows that

$$v \propto \sqrt{E_{da} I_d} \tag{12}$$

Inasmuch as it is necessary for the magnetic field to extinguish the arc, its best strength, β_g, is directly proportional to the electric field tending to maintain the arc. That is:

$$\beta_g \propto E_{da} \tag{13}$$

The number of ions to be removed from the gap is proportional to the current thru the arc, and hence to I_d. That is:

$$\beta_g \propto I_d \tag{14}$$

From equations 13 and 14:

$$\beta_g \propto E_{da} I_d \tag{15}$$

Hence from equations 10, 11, 12, and 15:

$$\beta_g = \frac{K E_{da} I_d}{\lambda \sqrt{E_{da} I_d}} = K \frac{\sqrt{E_{da} I_d}}{\lambda} \tag{16}$$

where K is a quantity *inversely proportional* to the specific molecular velocity of the gases surrounding the arc. For any particular gas it is a constant, the value of which is determined experimentally from observations involving the other quantities of equation 16. Its numerical value obviously depends upon the units employed.

Computation of K from the Chemical Analysis of the Gas

If kerosene $CH_3(CH_2)_8CH_3$ is used to supply the atmosphere for the arc it dissociates into $10C+11H_2$. The C precipitates and the arc is surrounded by an atmosphere of H_2 only.

If ethyl (grain) alcohol C_2H_5OH is used, it dissociates into CO_2+6H_2+3C. The carbon precipitates and the arc is surrounded by an atmosphere of H_2 diluted by CO_2.

The weights of equal volumes of the chamber gases may be computed from their molecular weights. Thus the

$$\frac{\text{density of chamber gas from kerosene}}{\text{density of chamber gas from ethyl alcohol}}=\frac{14}{56}=\frac{1}{4}$$

The velocities of the molecules of different gases, at the same temperature, are inversely proportional to the square roots of the densities of these gases. Hence the molecular velocity of the chamber gas with kerosene is twice that with ethyl alcohol.

The same method may be used for methyl (wood) alcohol, illuminating gas, and so on.

Thus the value of K, equation 16, for ethyl alcohol should be twice that for kerosene. This theory is proven from the following data.

Experimental Proof of Equation 16

These experimental data were taken at the United States Naval High Power Radio Station, Pearl Harbor, Hawaii, during the months of August and September, 1917. Experimental proof of equation 16 is given for powers up to 500 kilowatts thruout a wave length range of 4.1 to 16.1 kilometers. Values of K are derived for kerosene and ethyl alcohol.

In considering experimental data of this sort, it should be realized that there are many factors which render such a station unsuitable for tests requiring those features of unchanging conditions and ample time for observations which can only be obtained in the laboratory.

Antenna resistance R_s changes daily with the weather, for it is affected considerably by the surface condition of the antenna insulation. Furthermore, the field tuning is broad. These facts tend to scatter the observed points.

The lack of opportunity for long runs because of necessary routine work about the plant renders the obtaining of uniform chamber atmospheres on consecutive days practically impossible. This increases the scattering of the K determinations.

The reader will realize, therefore, that the data presented

in proving equation 16 are essentially those taken in the field and not in the laboratory with its attendant possible niceties of observation.

Figures 15 thru 22 (corresponding to plates 31, 52, 93, 134, 145, 156, 187, and 218) show the effect of variations in B_g and I_d with E_{da} held approximately constant thruout a range of wave lengths of from 4.5 to 16.1 kilometers.

These curves are practically equivalent (except for value of ordinate in amperes) to curves of I_s plotted against B_g, since I_s is proportional to I_d. The proportionality factor, ρ, varies with B_g for a given λ and E_{da}, but this variation is of no interest since we are interested only in the peak values of the curves, and these occur at the same B_g irrespective of which current is used in the scale of ordinates. At the peak of these curves, $\rho=\sqrt{2}$, and on each side of the peak it is greater than $\sqrt{2}$.

It will be noted that the points of maximum current fall approximately on a straight line thru the origin. This is in accordance with the theory used in the derivation of equation 14. Such lines have been drawn on all plates.

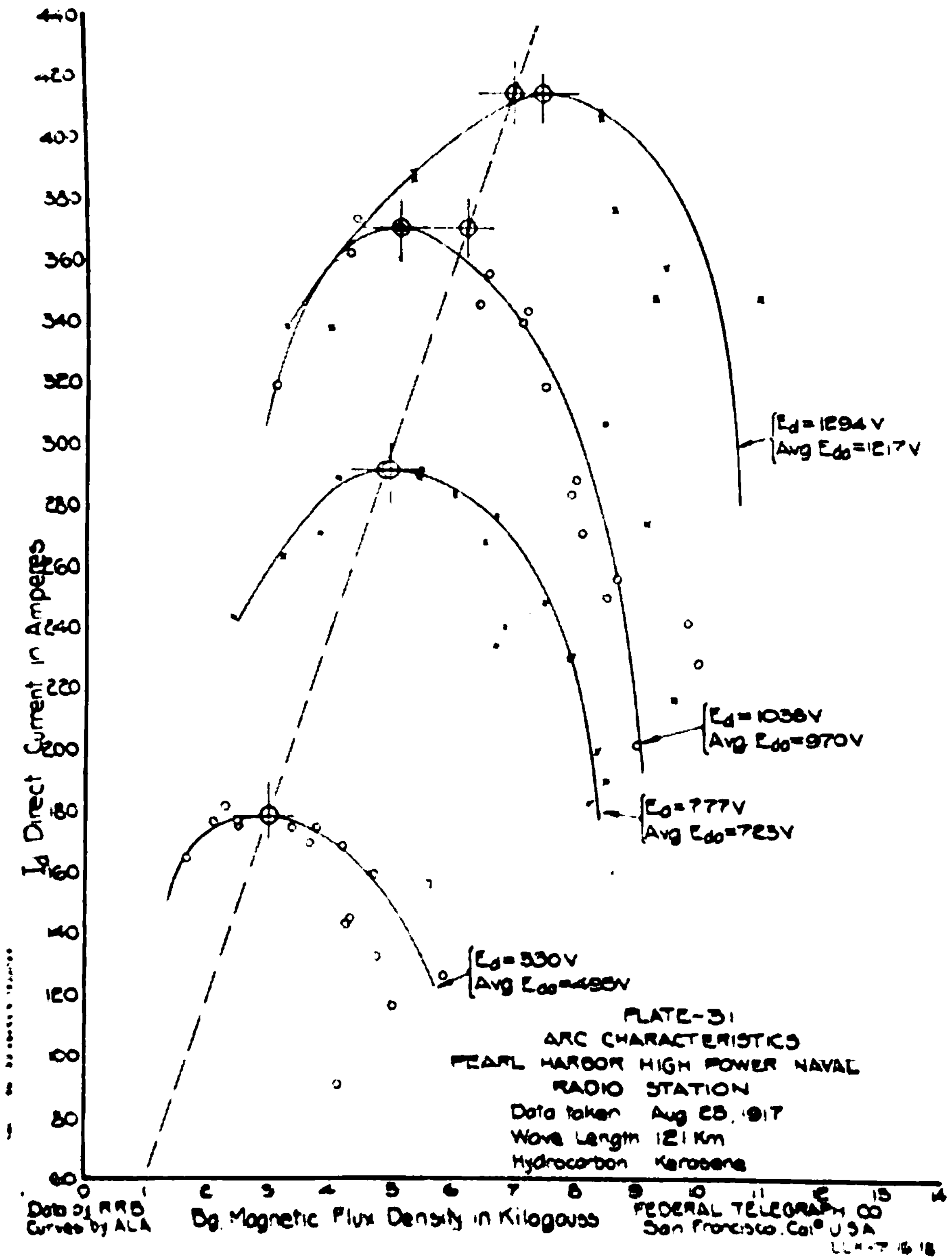


FIGURE 15

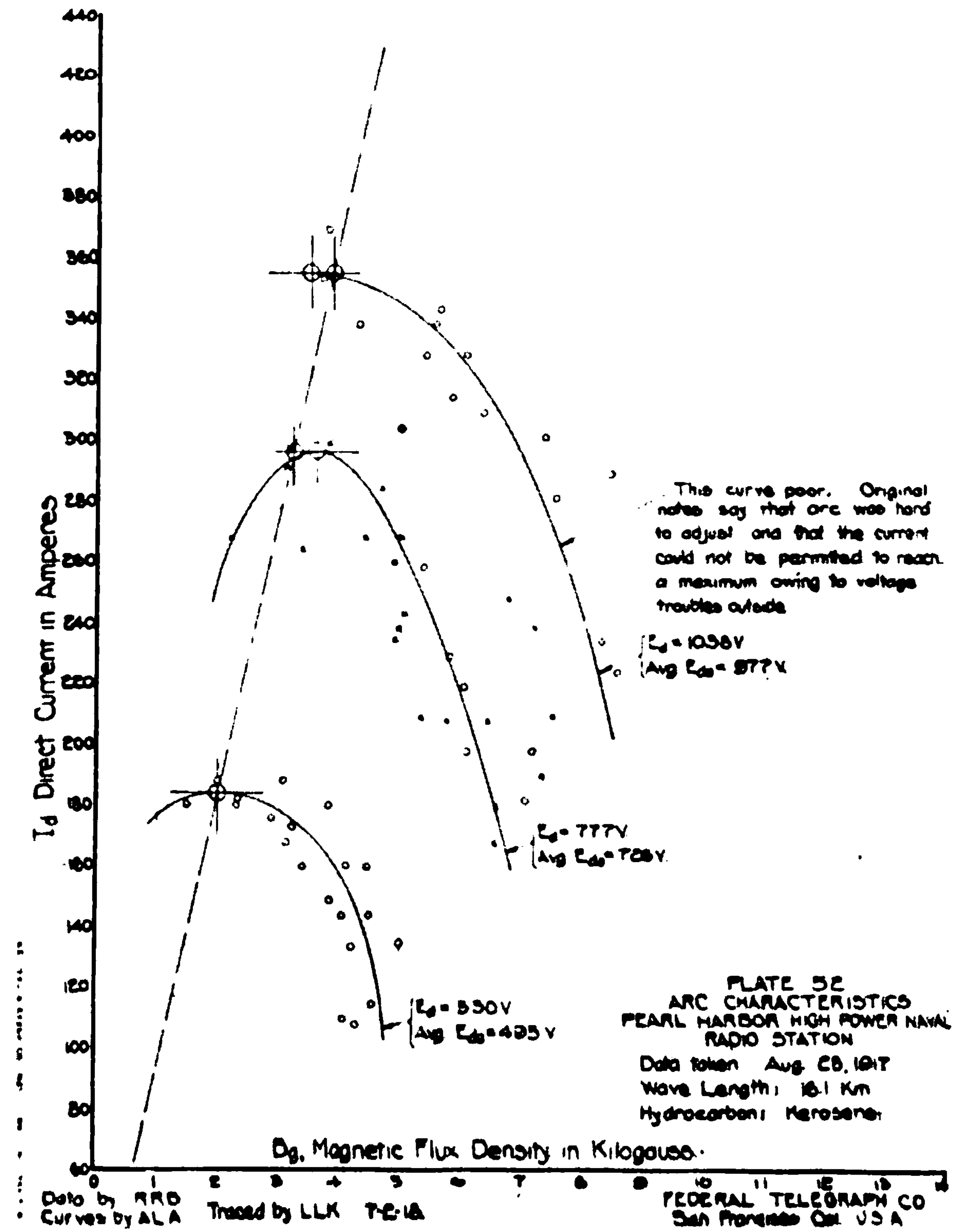


FIGURE 16

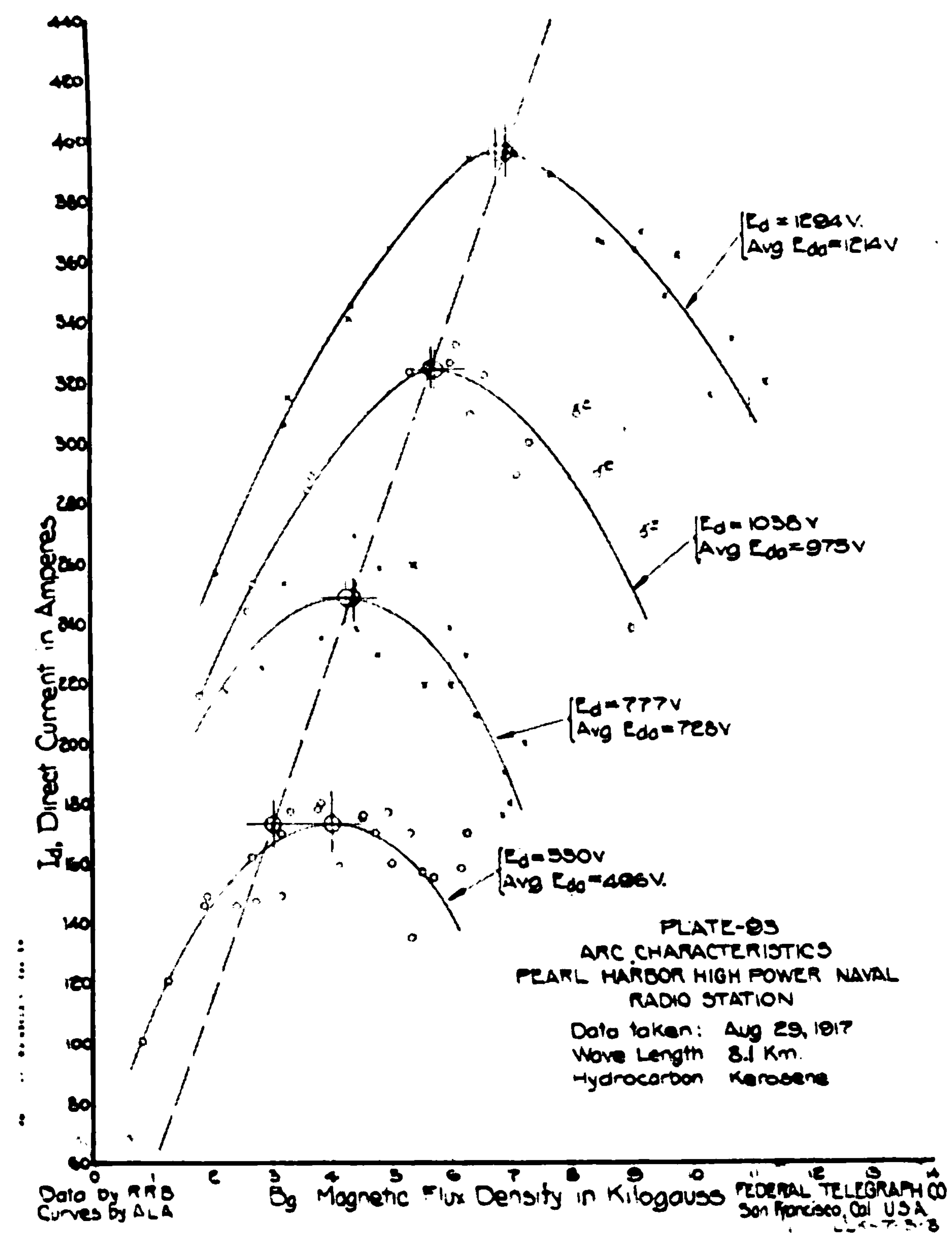


FIGURE 17

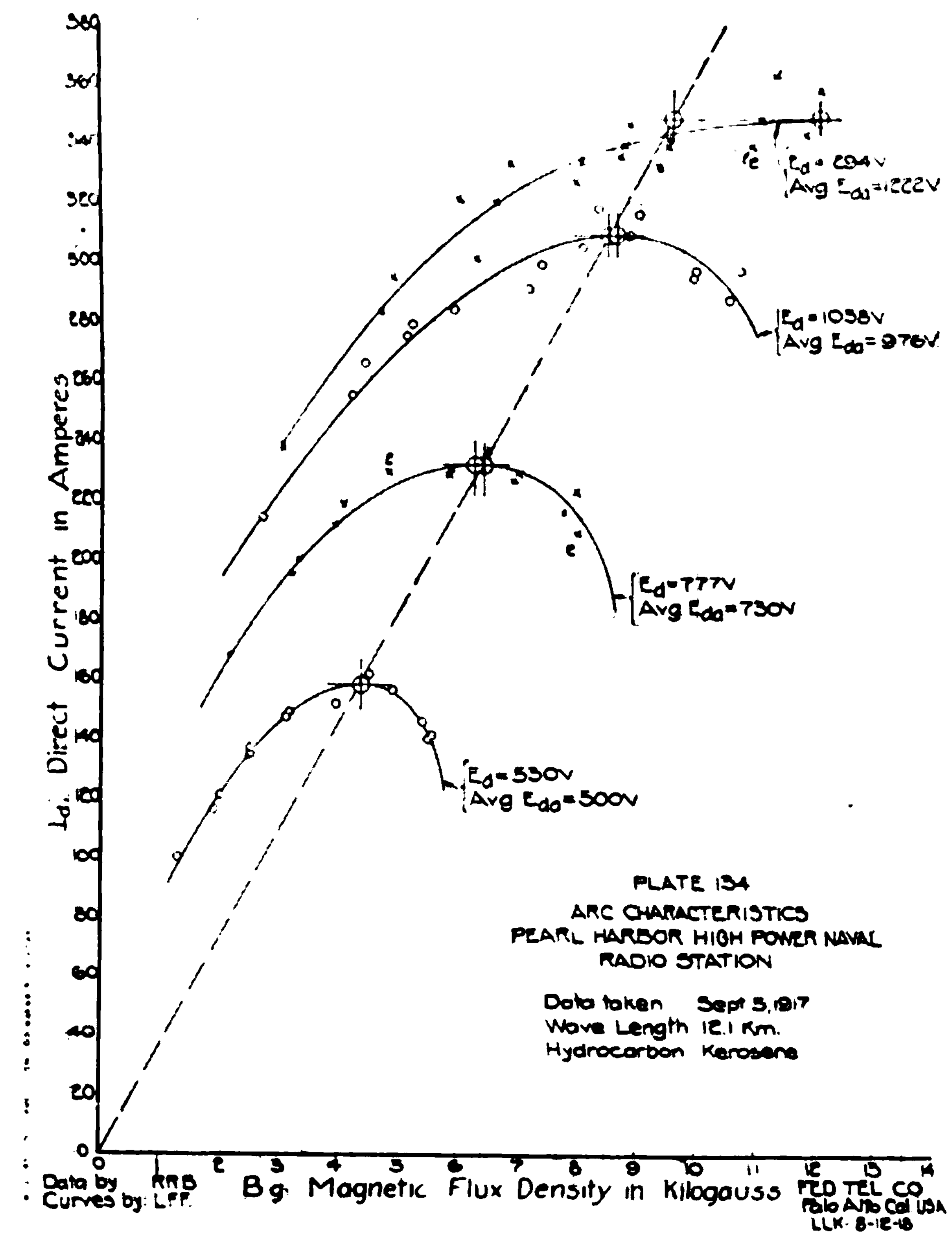


FIGURE 18

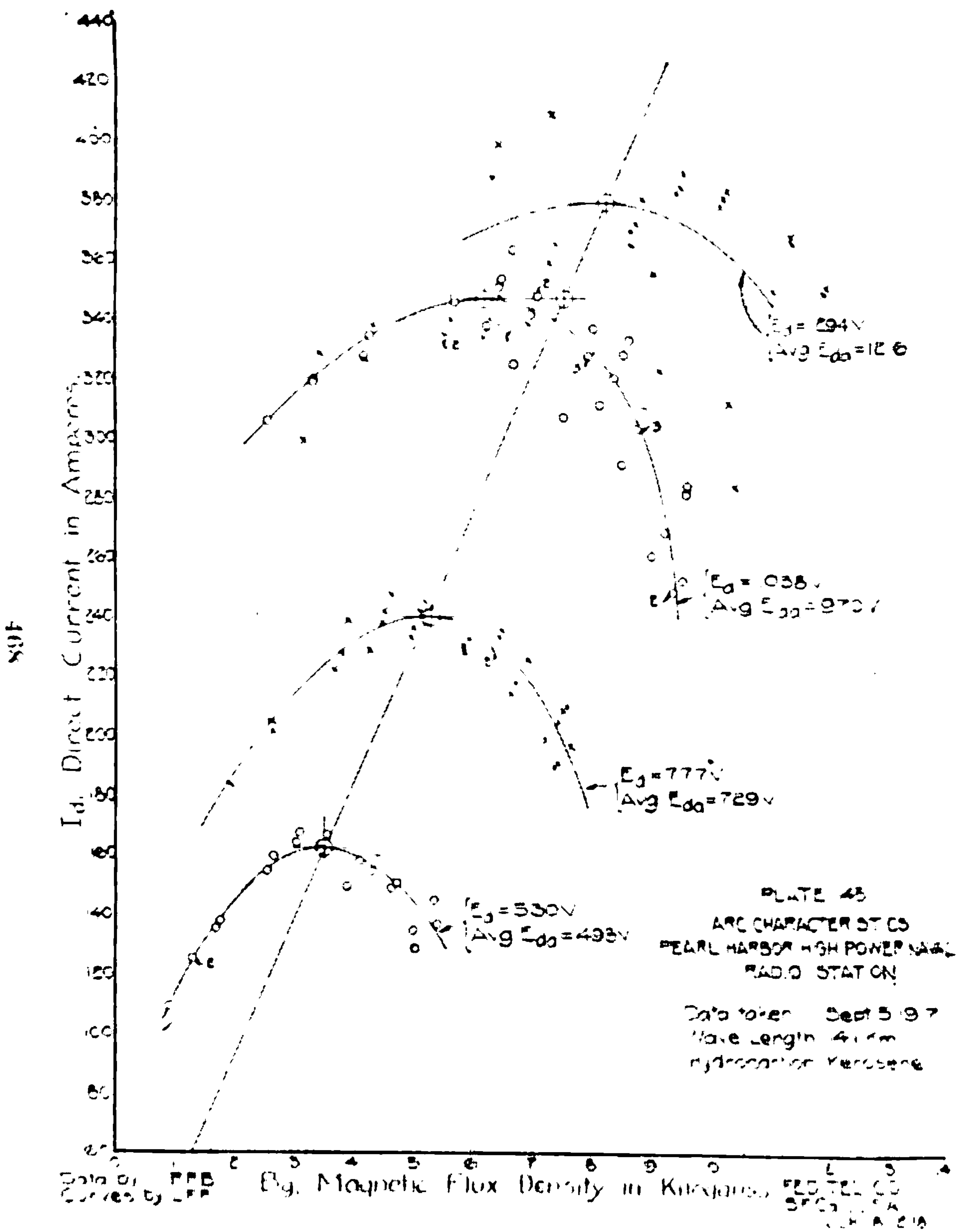


FIGURE 19

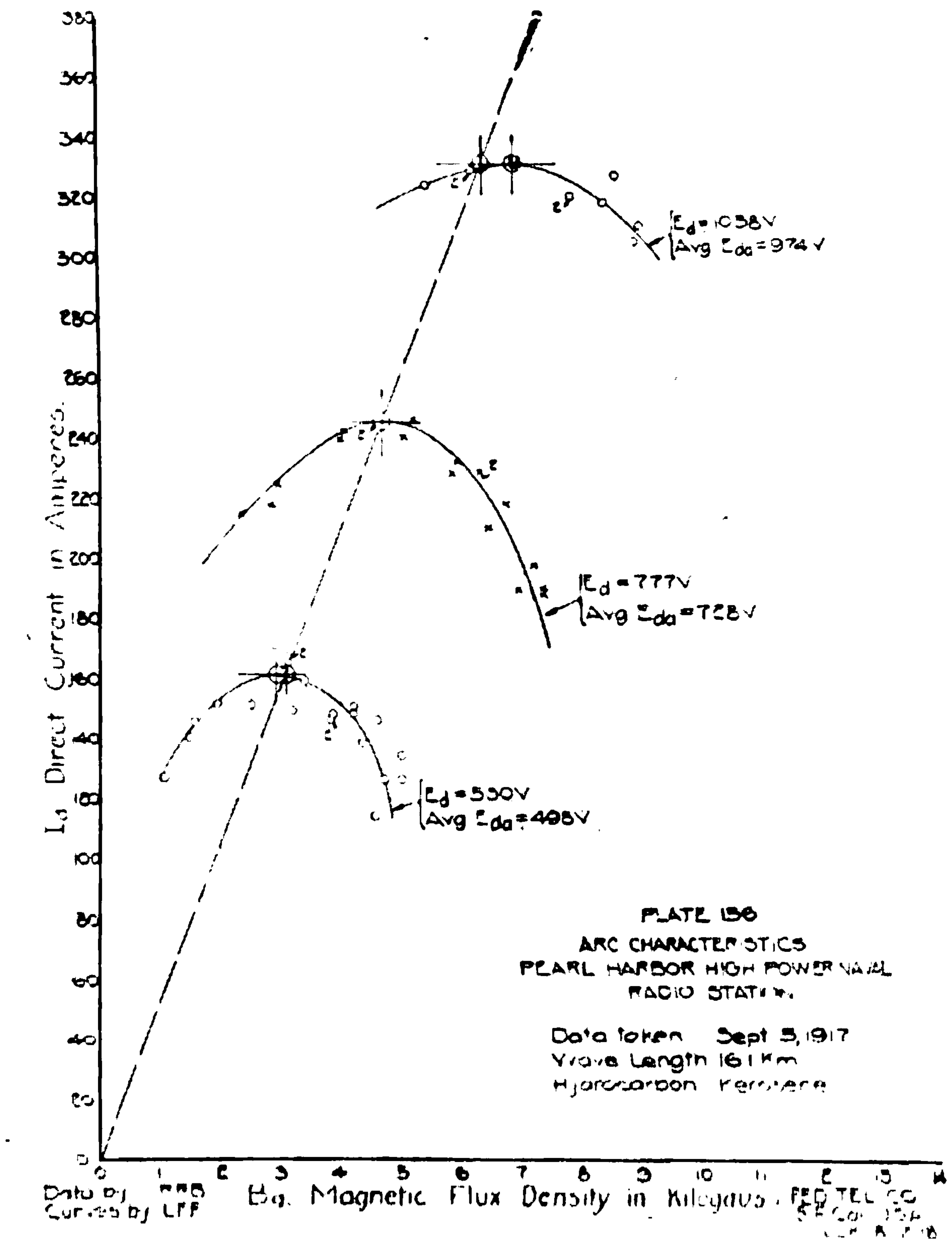


FIGURE 20

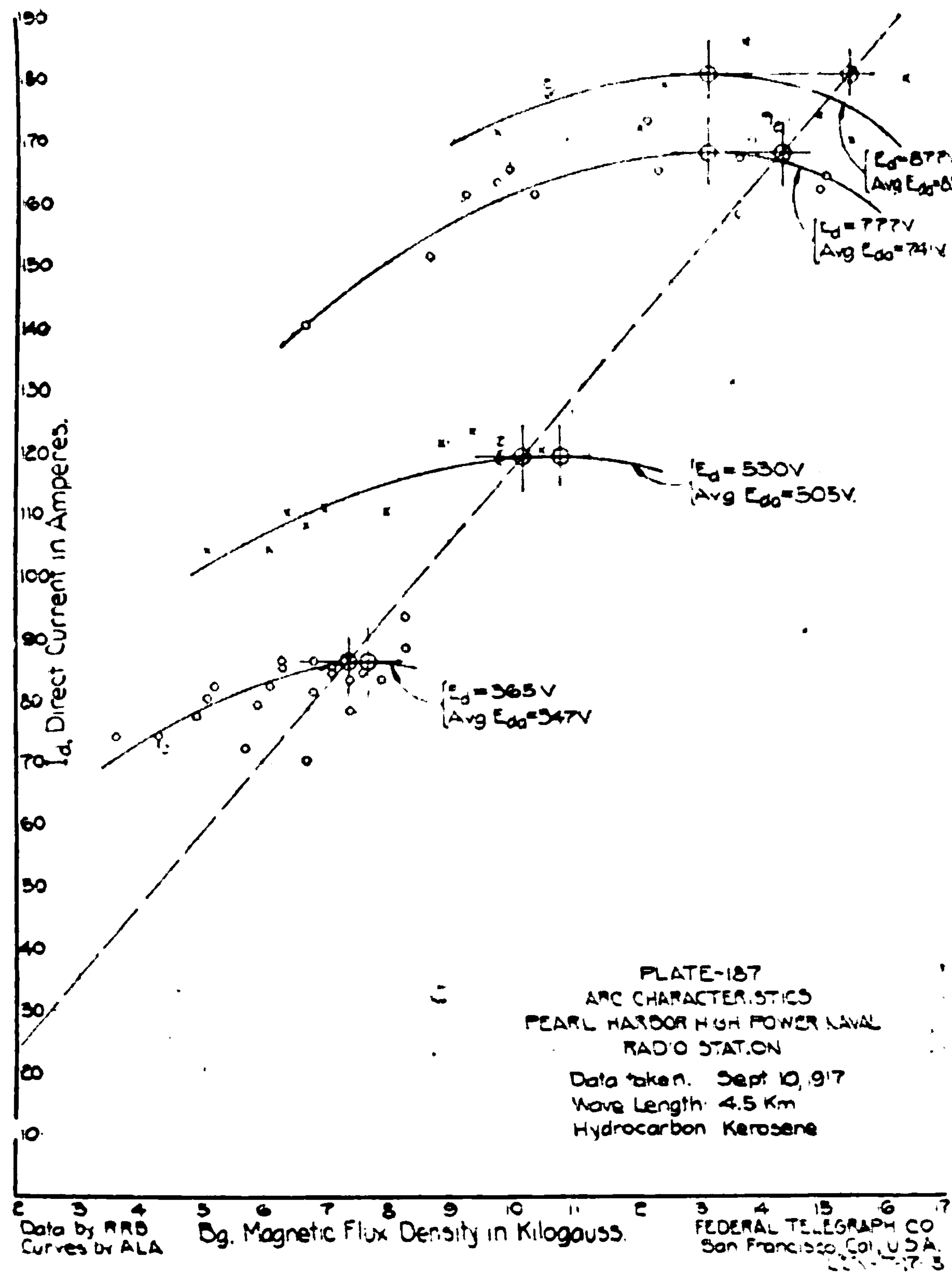


FIGURE 21

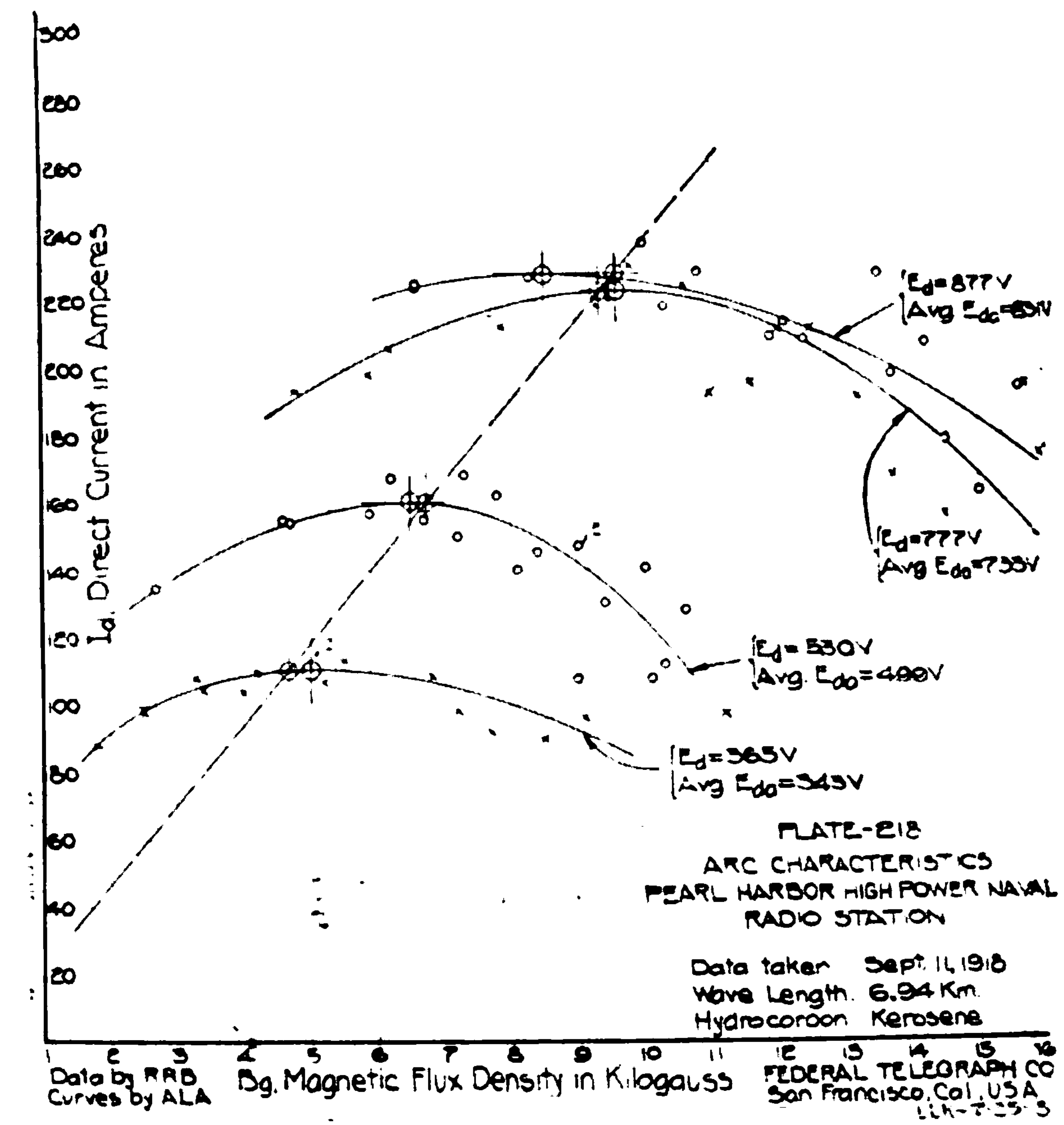


FIGURE 22

The peak values of I_d have been tabulated in Table 1 along with the value of β_g as determined by the intersection of the straight line thru the origin, with the current peak value.

TABLE 1

Figure Number	E_{da} Volts	I_d Amps.	$E_{da} I_d$ Kw. Input	β_g Kilogauss	$Tan\ a$ Slope	c Intercept	λ Wave Length	$c\lambda = K$
15	495	178	88.1	3.0	.532	.275	12.1	3.33
	723	292	211	4.85				
	970	371	360	6.4				
	1,217	415	505	7.4				
16	495	184	91.1	2	.504	.198	16.1	3.19
	726	296	215	3.2				
	977	355	347	3.87				
17	496	173	85.9	3.05	.477	.363	8.1	2.94
	728	248	180.5	4.35				
	975	324	316	5.7				
	1,214	396	481	7.0				
18	500	158	79	4.35	.497	.5	12.1	6.05
	730	232	169.5	6.4				
	976	310	302	8.5				
	1,222	348	425	9.55				
19	498	163	81.3	3.5	.510	.37	14.1	5.22
	729	240	175	5.2				
	970	347	336	7.5				
	1,216	379	461	8.2				
20	498	162	80.8	3.05	.499	.35	16.1	5.64
	728	246	179	4.75				
	974	332	323	6.25				
21	347	86	29.8	7.4	.464	1.55	4.5	6.98
	505	119	60	10.2				
	741	168	124.5	14.4				
	838	180.5	151.2	15.5				
22	343	110	37.7	4.7	.466	.86	6.94	5.97
	499	160	79.9	6.75				
	733	222	163	9.35	Aver. .494		Aver.	4.91
	831	228	189.5	9.6				

The kilowatt input has been computed and tabulated, and on Figure 23 are logarithmic graphs of the relation between $E_{da} I_d$ and β_g. The slope of these lines is tabulated in Table 1, and averages 0.494. The theoretical value is 0.500 as indicated by equation 16.

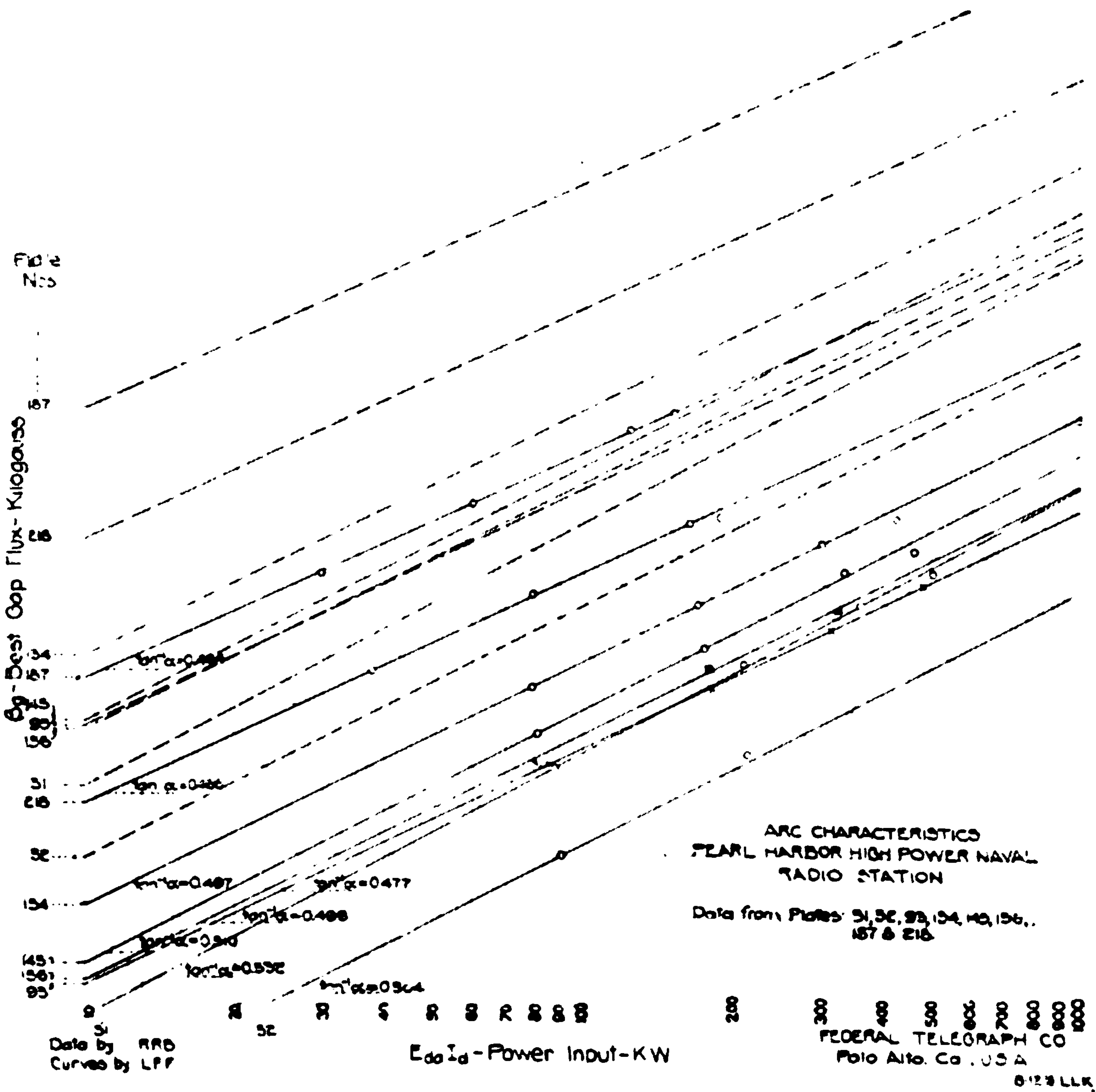


FIGURE 23

Dotted lines show the intercepts on the unit axis. These have been tabulated also. According to equation 10, these intercepts should be inversely proportional to wave length. Therefore the product $c\lambda$ has been tabulated, and the average 4.91 obtained.

We have thus obtained an experimental check upon equation 16, and determined the empirical value of

$$K = 4.91$$

when kerosene is used to supply the arc atmosphere and

E_{da} is expressed in kilovolts
I_d is expressed in amperes
λ is expressed in kilometers
β_g is expressed in kilogauss.

Figures 24 thru 30 (corresponding to plates 231, 242, 253, 264, 285, 296, and 307) repeat the foregoing when ethyl alcohol is used. Figure 31 and Table 2 give the results.

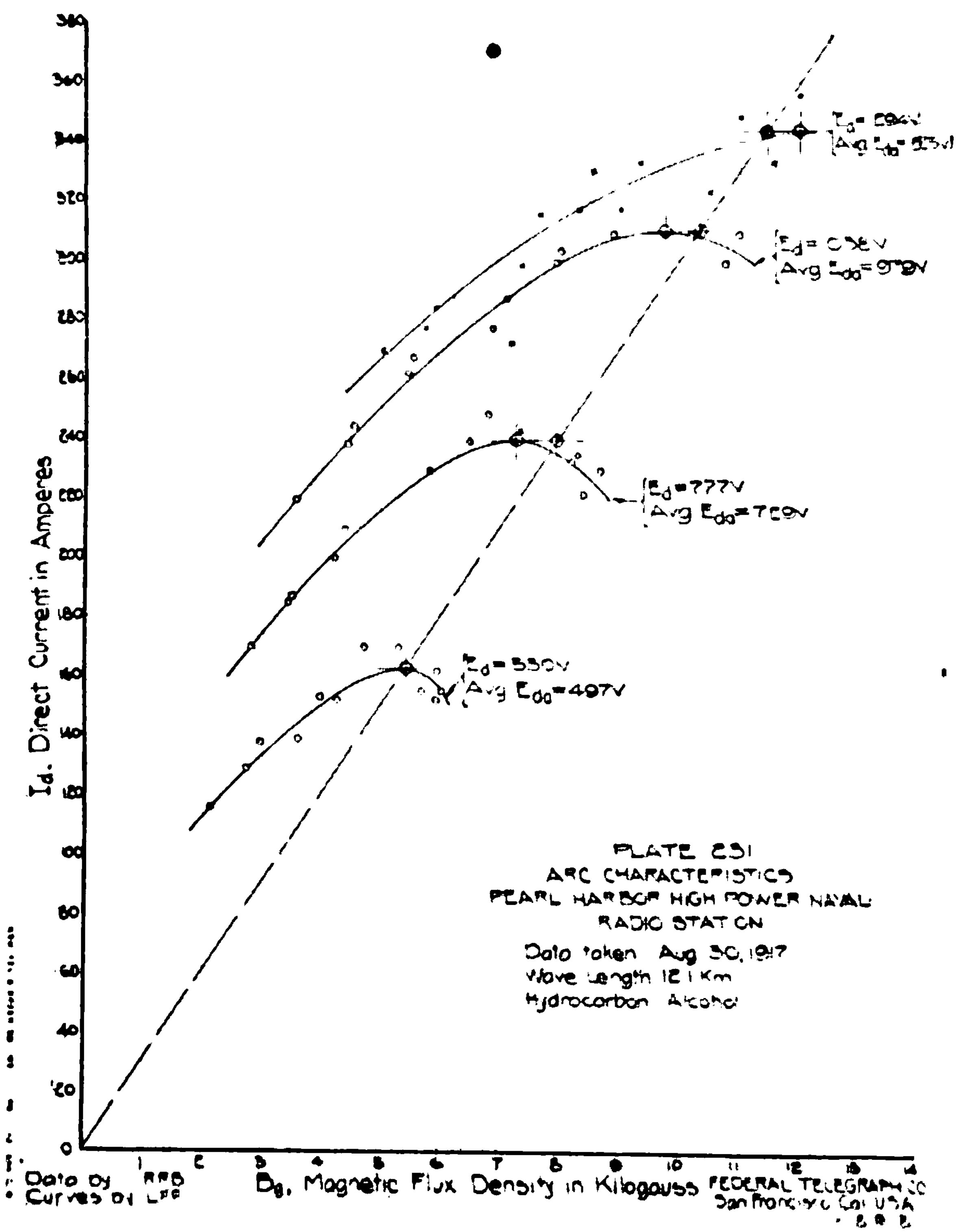


FIGURE 24

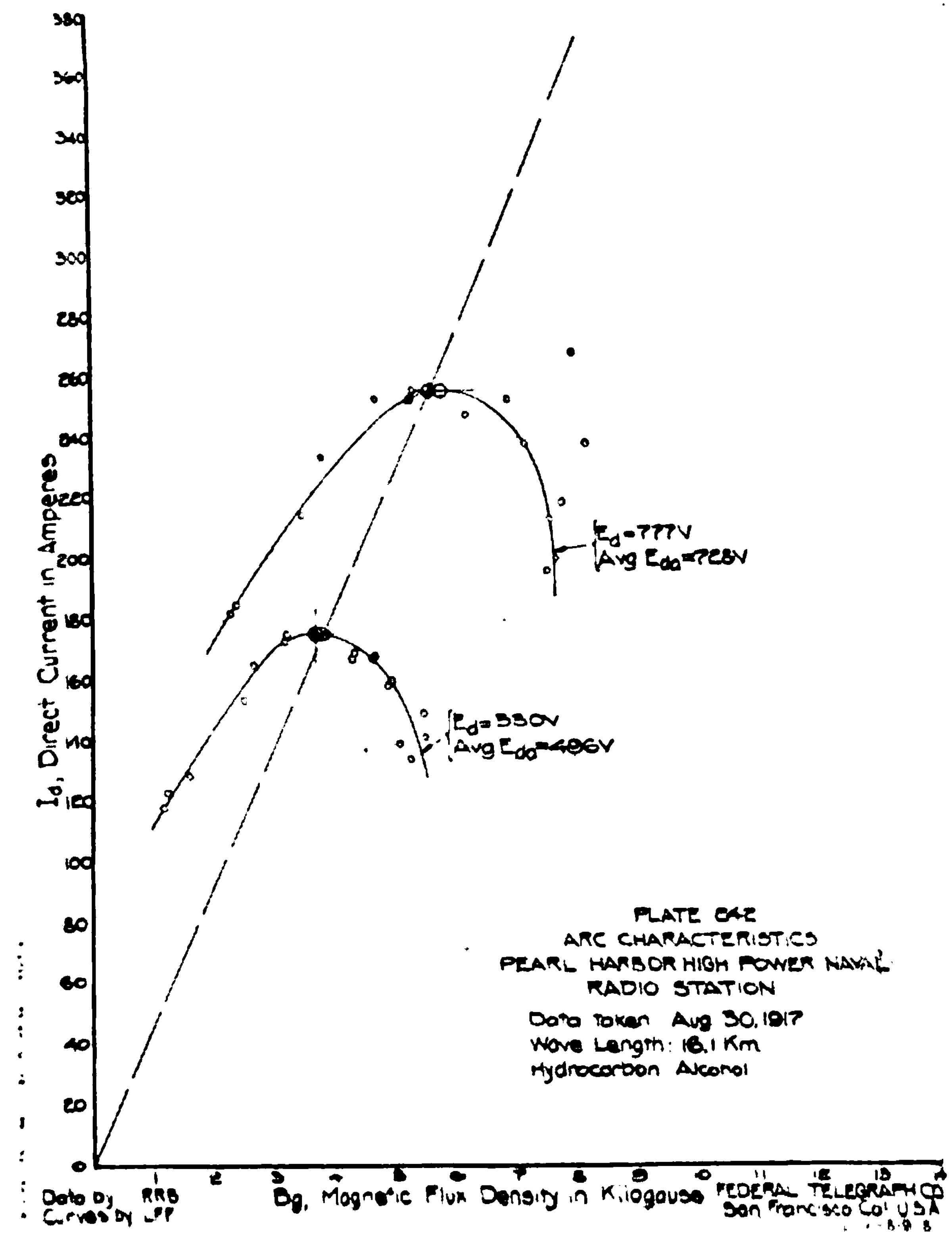


FIGURE 25

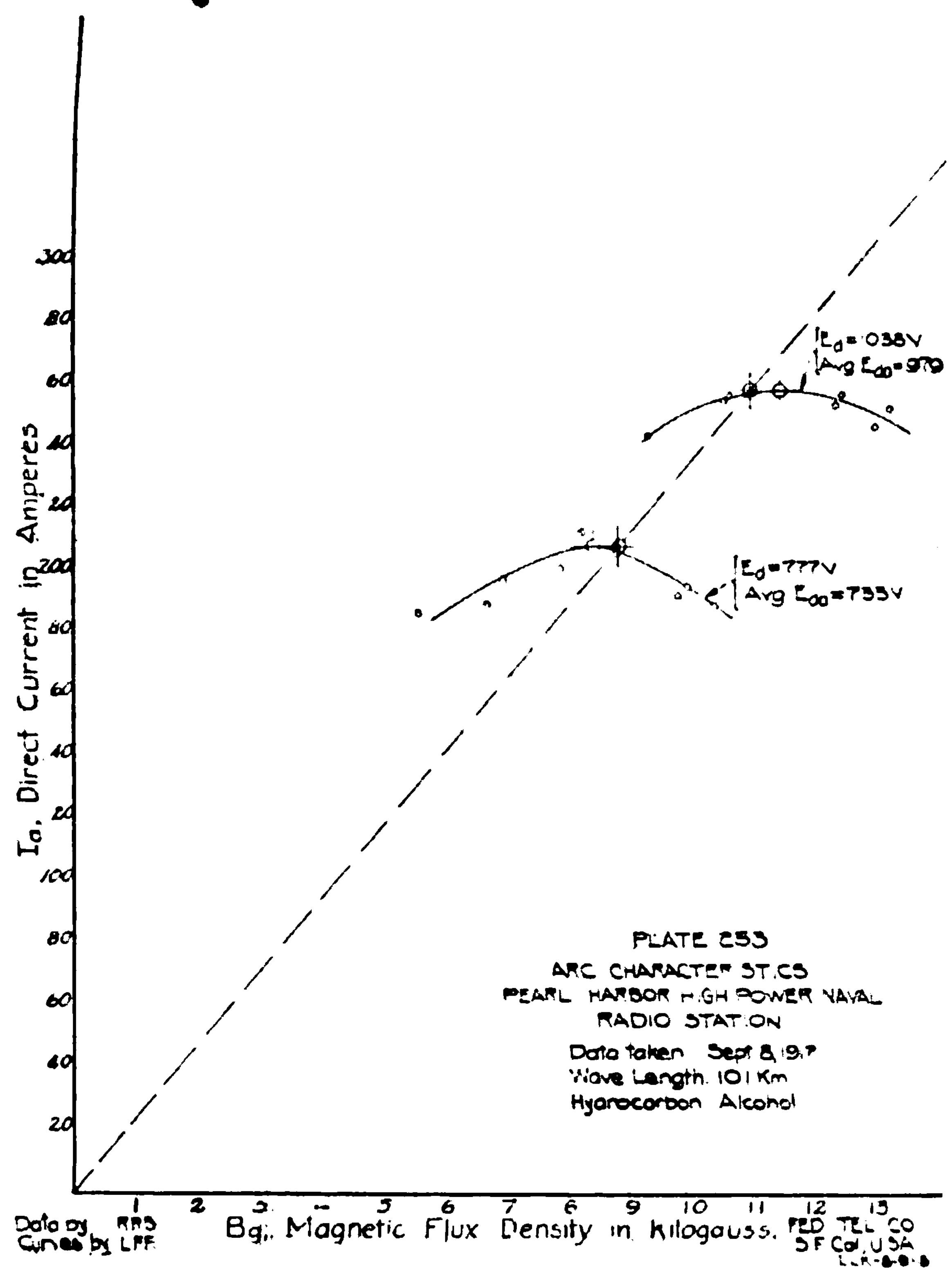


FIGURE 26

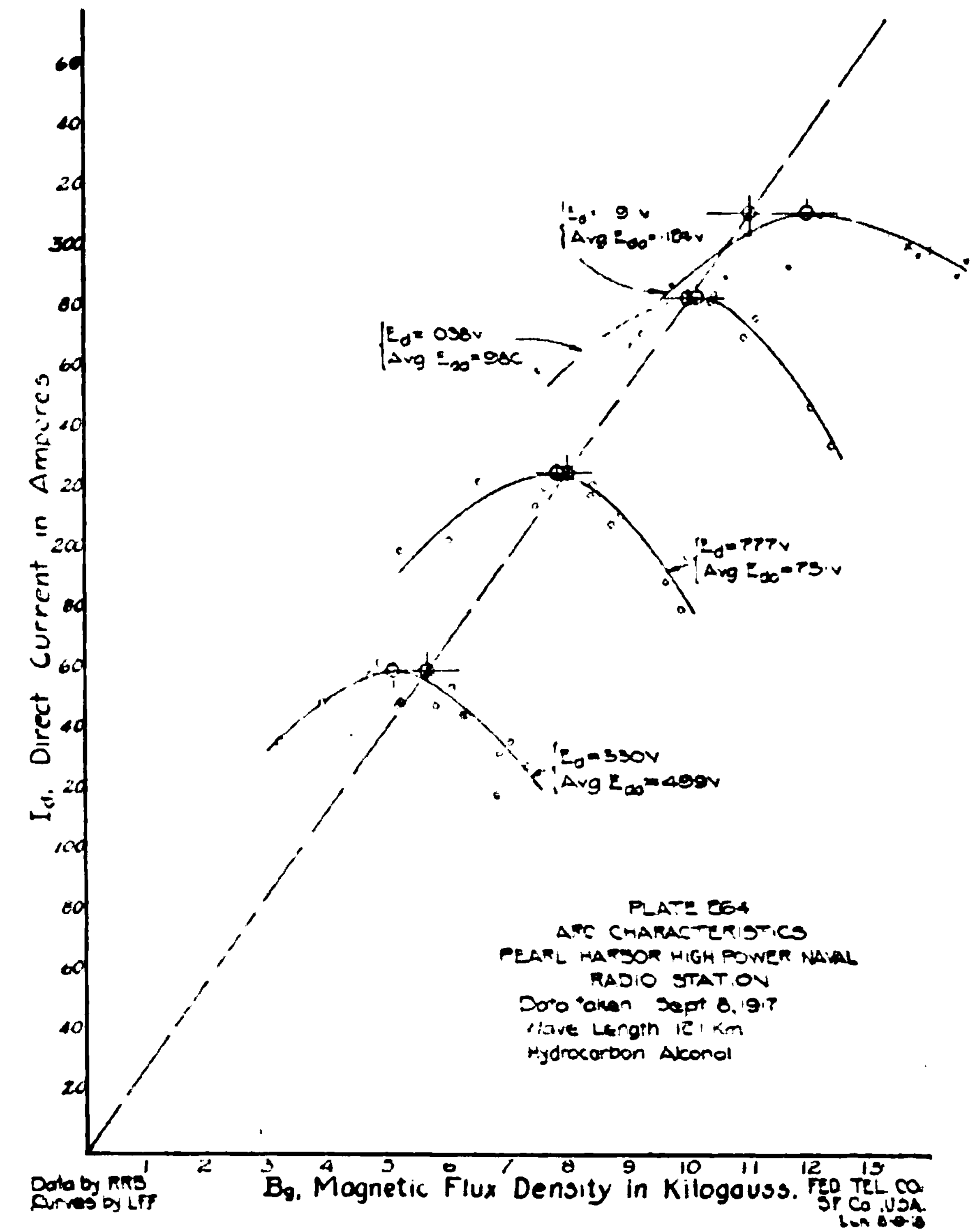


FIGURE 27

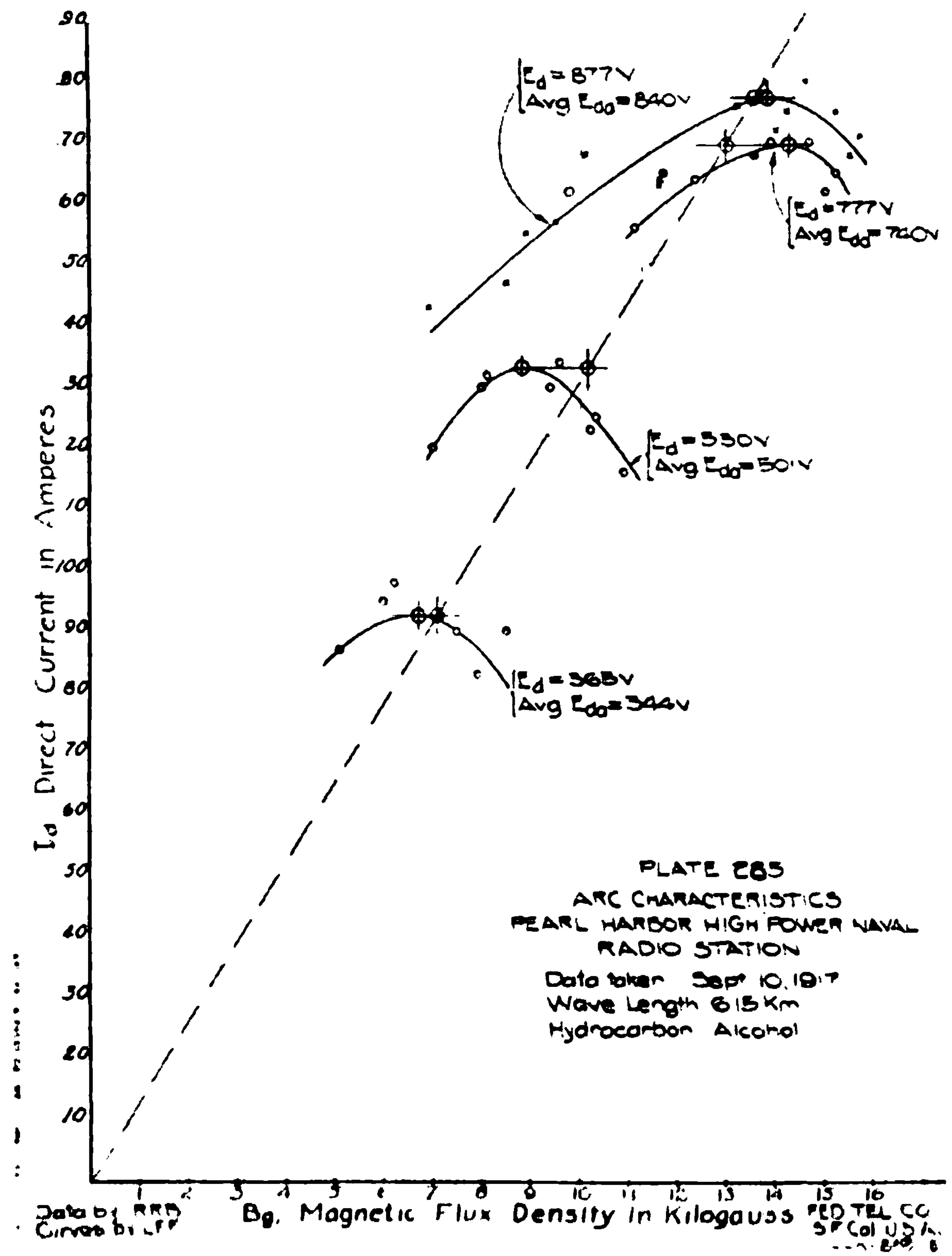


FIGURE 28

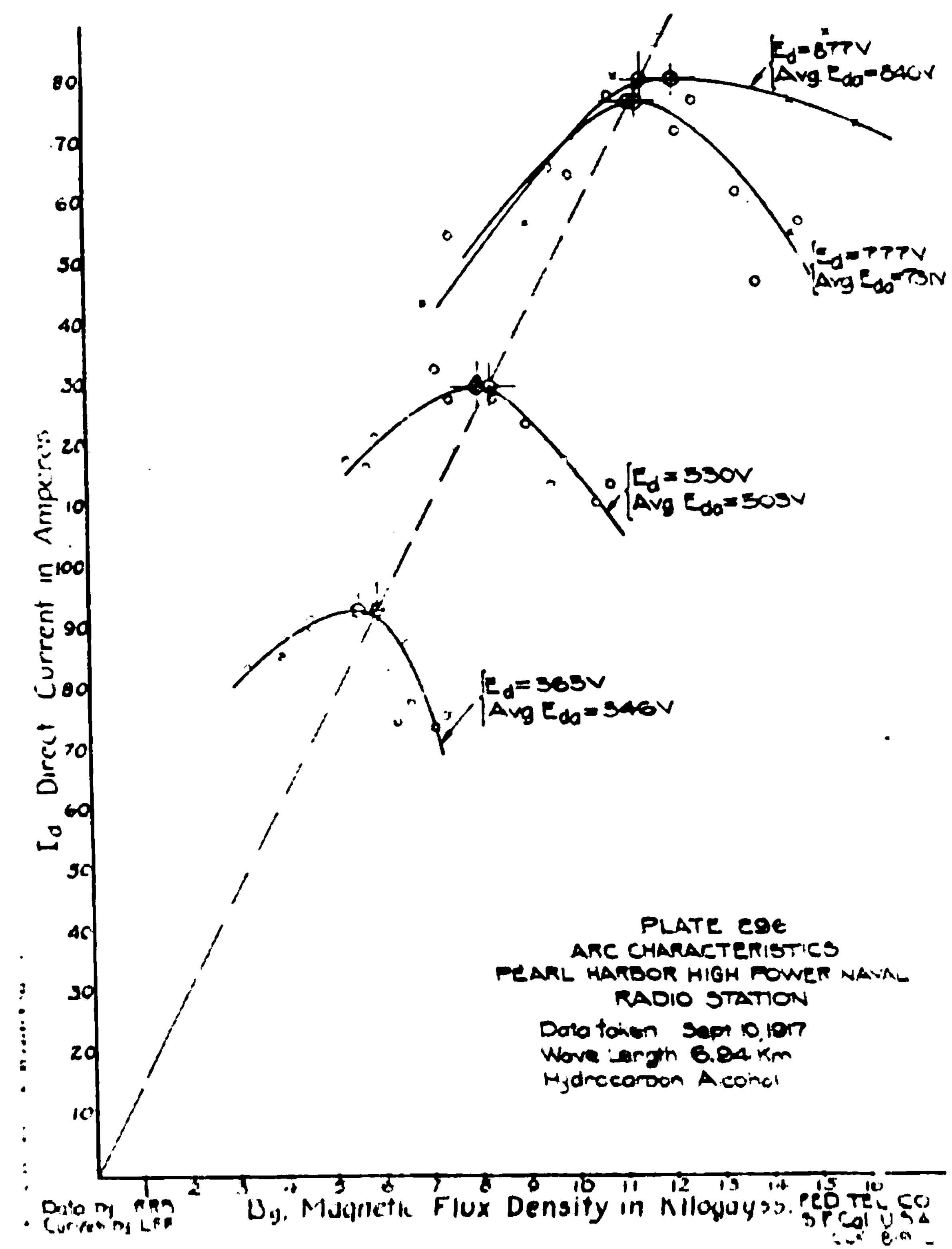


FIGURE 29

The average slope is 0.483, and the value of

$$K = 8.32$$

Our theory, given above, showed that the value of K for ethyl alcohol should be twice that for kerosene.

$$2 \times 4.91 = 9.82 \text{ (as against 8.32)}$$

We therefore check our theory within 18 per cent. It is to be remembered that this theory is based upon chemically pure liquid hydrocarbons of the molecular make-up indicated, and that in these tests commercial kerosene and grain alcohol were used. The impurities or variations in the molecular make-up of these would in some degree affect the results.

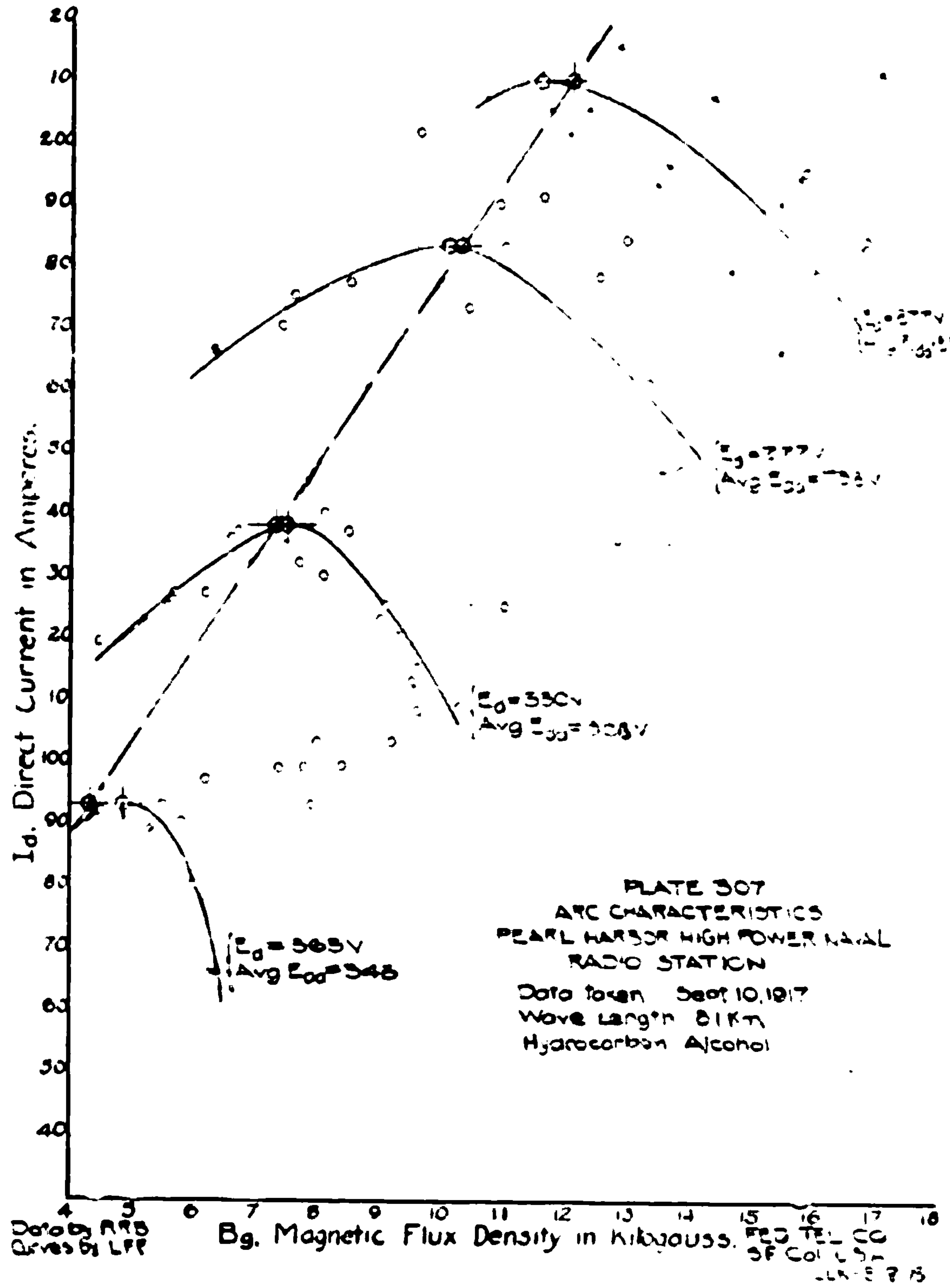


FIGURE 30

Any theory which assumes ideal conditions can only indicate the general trend of the phenomena which will take place under practical conditions.

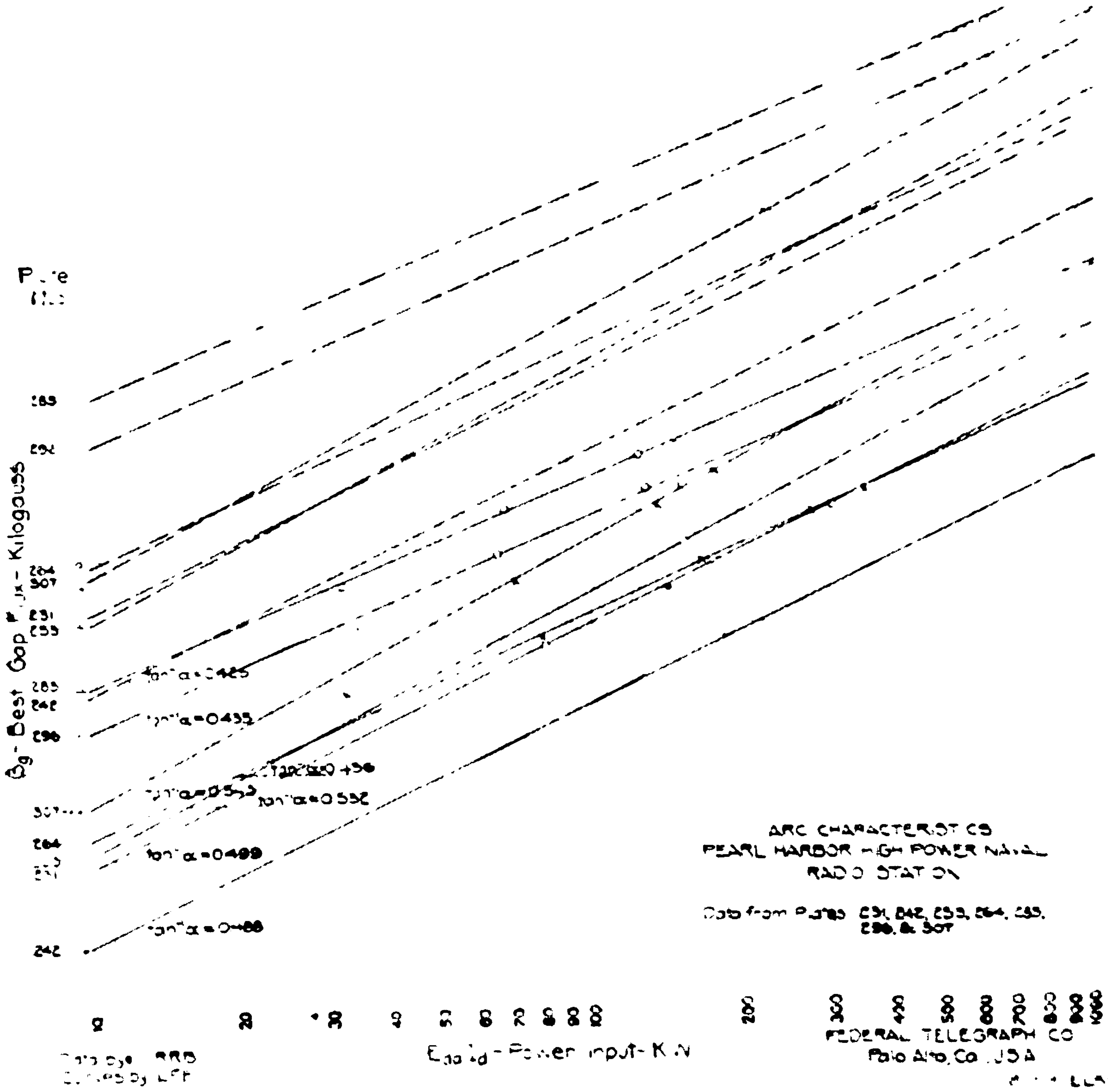


FIGURE 31

In view of the disturbing factors which are known to affect arc performance, such as the impurities in the hydrocarbon above mentioned, and especially those unavoidable variations in the density of the chamber atmosphere previously enumerated, it is felt that an agreement even as close as 18 per cent. establishes the soundness of our fundamental conceptions of arc operation.

While variations in chamber atmospheric density affect the position of the straight lines on the logarithmic sheets and hence the value of K so obtained, they do not affect the slope of these lines, provided the rate of hydrocarbon flow is held constant thruout any one run. For this reason the determination of the exponent of the $E_{da} I_d$ product is checked rather closely, that is, within 1.25 per cent. in the case of kerosene and 3.5 per cent. in the case of ethyl alcohol.

TABLE 2

Figure Number	E_{da} Volts	I_d Amps.	$E_{da} I_d$ Kw. Input	B_y Kilogauss	Tan α Slope	c Intercept	λ Wave Length	$c\lambda = K$
24	497	162	80.5	5.4	.499	.62	12.1	7.5
	729	239	174.2	8.0				
	979	310	303.5	10.35				
	1,225	344	420.2	11.45				
25	496	175	86.9	3.85	.488	.46	16.1	7.4
	728	256	186.5	5.6				
26	733	209	153.0	8.7	.532	.60	10.1	6.06
	979	259	253.5	11.3				
27	499	160	79.9	5.6	.456	.78	12.1	9.44
	731	226	165.2	7.95				
	980	284	278.5	10.0				
	1,124	312	351.0	11.0				
28	344	92.5	31.8	7.19	.425	1.68	6.15	10.33
	501	133	66.6	10.13				
	740	169.5	125.5	13.0				
	840	177	148.7	13.5				
29	346	97	33.6	5.88	.435	1.35	6.94	9.37
	503	130	65.5	8.25				
	731	177	129.4	11.13				
	840	180	151.2	11.38				
30	345	94	32.4	4.35	.543	.74	8.1	6.0
	503	139	70.0	7.3				
	738	184	135.8	10.3	Aver. .483		Aver.	8.32
	832	211	175.5	12.05				

In taking these data, it was customary to use alcohol profusely because there was little soot deposited in the chamber, and this practice assured the most uniform results. When kerosene was used, the soot deposit became excessively heavy, and it was therefore customary for convenience in arc operation to throttle down the flow somewhat. This reduced the resultant effective molecular velocity of the chamber gases because they were diluted by O_2 and N_2 from the outside atmosphere which always tends to be drawn thru small leaks into the swirling mass of hot gases within the arc chamber when the liquid hydrocarbon is supplied in insufficient quantity.

It is evident from the foregoing that the value of K determined for kerosene is high (considering the ideal conditions for which equation 16 was derived), and in view of the practical difficulties of taking experimental data of this sort, some of which have already been enumerated, we may generalize some-

what, and are of the opinion that $K = 4.25$ for kerosene and 8.5 for ethyl alcohol are better empirical determinations of this constant. Equation 16 becomes:

$$\beta_g = \frac{4.25 \sqrt{E_{da} I_d}}{\lambda} \text{ for kerosene} \tag{17}$$

and

$$\beta_g = \frac{8.50 \sqrt{E_{da} I_d}}{\lambda} \text{ for ethyl alcohol} \tag{18}$$

General observation and data collected from various sources indicate that, for ordinary Pacific Coast illuminating gas, the value of K lies about midway between that for kerosene and ethyl alcohol.

A consideration of the chemical make-up of methyl alcohol shows that it should require a somewhat higher flux than ethyl alcohol. This is borne out by general experience in arc work.

The Magnetic Circuit

General

In our treatment of the arc design problem thus far, we have covered the theoretical field of arc performance and have shown the relationship between the various voltages and currents involved from theoretical considerations backed by experimental evidence.

We have also derived the so-called "Flux Formula" (Equation 16) from theory, and have proven it experimentally by data taken at the Pearl Harbor High Power Naval Radio Station.

We shall now consider the theories involved in the design of the magnetic circuit and shall follow these by experimental data bearing specifically upon the affect of variations in magnetic circuit design on the gap flux, B_g.

Until the high power arc converter became a necessity in radio telegraphy, this specific problem had not been met in practical design work, except in the case of a few relatively small electro-magnets which had been built for various laboratories thruout the world.

The necessity of building electro-magnets weighing 65 tons (59,100 kg.) in the case of the 500 kw. arc converters and 80 tons (72,700 kg.) in the case of the 1,000 kw. units made it essential that knowledge of these matters be accumulated in a form suitable for design work.

The Ideal Circuit

Figure 32 shows two magnet poles "N" and "S" located under ideal conditions as developed in past literature—that is, within a long solenoid, so that at the magnetic air gap the magneto-motive force (abbreviated "mmf.") is uniformly distributed and is parallel to the axis of the poles. If it is desired to

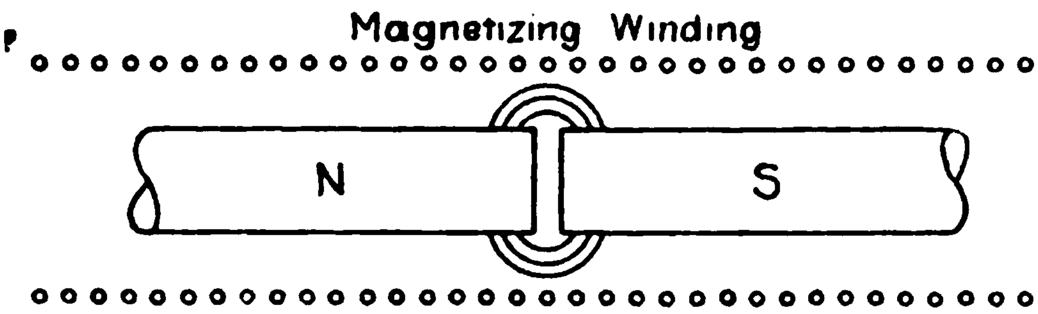


Figure 32

obtain a maximum flux density in the air gap for a given applied mmf., it is obvious that this cannot be obtained if the full diameter of the poles is continued up to the air gap, because a certain percentage of the flux will leak around the gap as shown, and the flux density in the main poles will exceed that in the pole tip faces adjacent to the gap. Hence premature saturation will occur back in the body of the main poles.

If the poles are shaped as in Figure 33, premature saturation of the main pole is eliminated, but an analogous condition exists in the pole tips—that is, premature saturation occurs at some point in the pole tip between its base, where it joins the main pole, and its face, adjacent to the air gap.

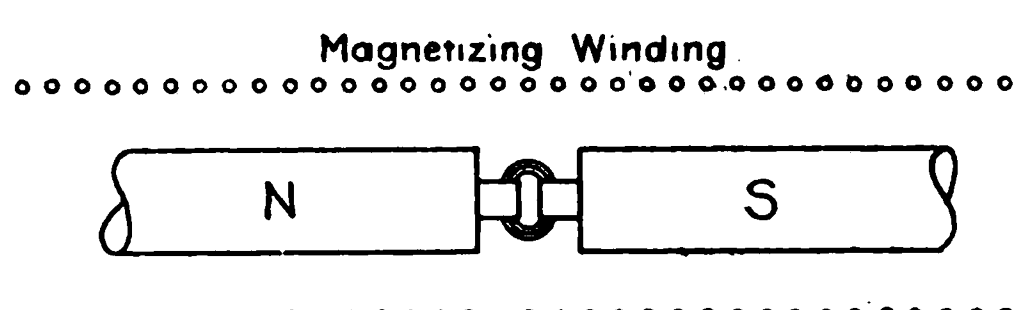


Figure 33

It is obvious from the foregoing that if the pole tips are made of a shape somewhat between that of Figures 32 and 33—for example, truncated cones as shown in Figure 34, a maximum flux will be obtained in the air gap for a given applied mmf. *If the poles are made of the best possible shape, equal flux densities*

will exist at all points, and hence no part will take more than its share of the applied mmf.

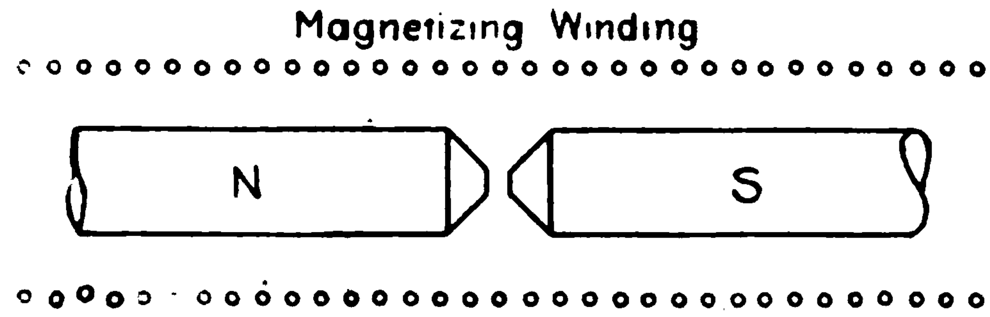


FIGURE 34

In his "Magnetic Induction in Iron and Other Metals," Ewing shows by analytical treatment that the best pole shape is that of a truncated cone having the angle α, Figure 35, $54° 44'$. His method of treatment is as follows:

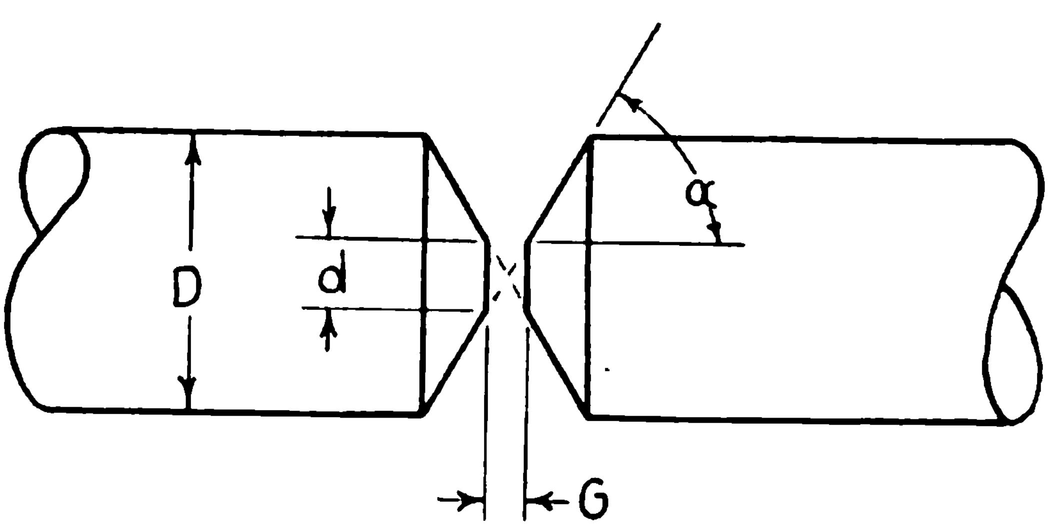


FIGURE 35

The assumption is made that the distribution of magnetization is uniform thruout the cross section of the magnet poles—that is, that the flux lies parallel to the axis of the poles.

The magnetic force in the space between the pole tips is composed of two parts, (1) the magnetic force due directly to the current in the field coils, and (2) that due to the internal or molecular mmfs. induced by the applied mmf. The first of these forms a small part of the whole, and since its distribution is nearly uniform, it becomes a negligible factor.

The pole faces are considered as being made up of a series of coaxial circular rings in planes normal to the axis of the poles. If the induced magnetization of one of these rings is represented by "J" the magnetic force "F"[1], due to it at a point "O" (Figure

[1] Using Ewing's terminology.

36) in the axis at a distance "x" from the plane of the ring, is, according to Coulomb's law, given by the equation:

$$F = \frac{J}{l^2} \cdot \cos a, \text{ or by}$$

$$F = \frac{Jx}{l^3} \quad \text{since } \cos a = \frac{x}{l}$$

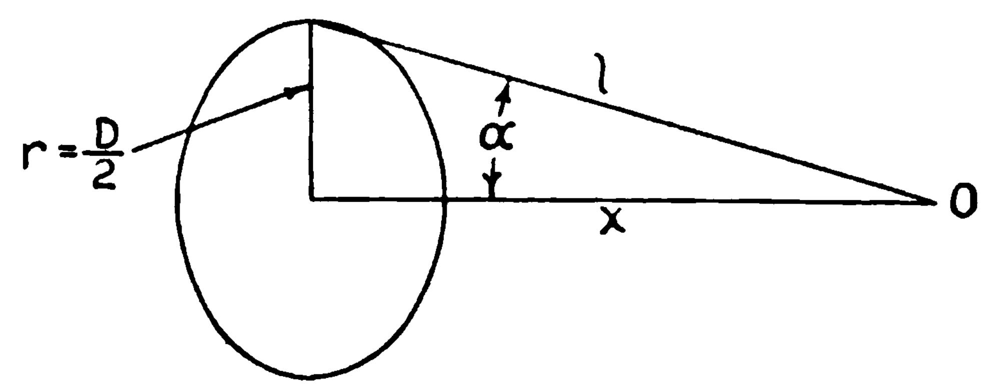


FIGURE 36

Obviously F will be a maximum when $\frac{dF}{dx} = 0$, which occurs when $x = \frac{r}{\sqrt{2}}$; $\tan a = \sqrt{2}$, or $a = 54° 44'$.

Weiss[2] has given experimental data showing that the best pole tip shape is not the uniform cone just considered but one made of a multiple number of angles, which are equivalent to a curved surface. This deviation from the pole tip shape which Ewing's mathematical treatment of the problem indicated as best, is probably due to the following facts:

(1) The fundamental requirement is that premature saturation shall not occur in any part of the magnetic circuit, that is, all parts must reach their limiting value at the same time.

(2) Such a condition was assumed by Ewing in his mathematical analysis, upon the basis of the flux in all parts of the pole tip lying parallel to the axis of the poles.

(3) But experimental data show that the flux does not lie parallel to the polar axis with uniformly conical tips. On the contrary, it concentrates in the portion next to the air gap.

(4) Therefore, in order to fulfil the basic requirement of (1), it is necessary to approach the gap more slowly and thus by

[2] For various discussions of pole tip angles, see
Weiss, "Journal de Physique," volume 6, page 353, 1907.
DeBois, "Ann. d. Phys.", volume 37, page 1268, 1913.
Ewing, former citation.
Cotton, "Revue Générale des Sciences Pures et Appliquées," volume 25, numbers 13 and 14, 1914.

leakage prevent too high a percentage of the main pole flux from flowing thru the tip faces adjacent to the gap.

Figure 37 shows a pole tip of this shape. It is relatively long as compared with single angle cones of approximately 55° and since, at the flux densities used in arc converters, the percentage of gain in B_g due to its use is small, it is not a desirable shape, because such a long-nosed pole materially increases the height of the magnetic circuit yoke. In arcs of 500 to 1,000 kw.

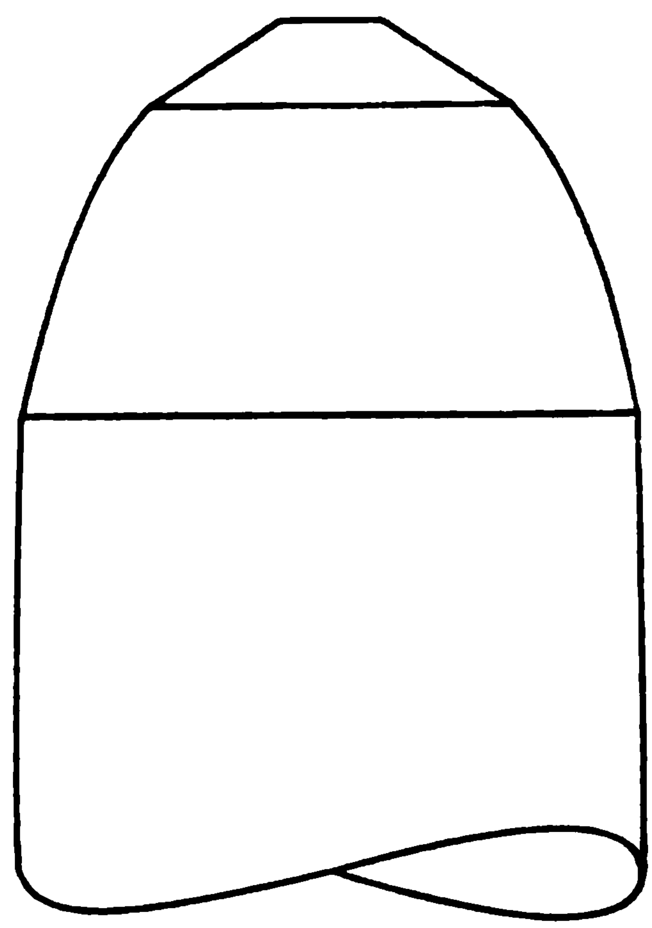


FIGURE 37

capacity, such an increase adds many tons to the weight of the unit, while the resultant decrease in necessary magnetizing kilowatts is very small. The time and cost of machining this complicated surface is another reason for not considering it in large arcs.

Figures 38 and 39 of a 1,000-kw. arc converter show why weight is added rapidly if the height of the yoke is increased.

The greatest flux density will be produced in the air gap when the pole pieces are saturated so that the intensity of induced magnetism J reaches its limiting value in all parts of the metal. Thus, for truncated cones (Figure 35) the surface density is $J \sin a$; and, employing the theory of Ewing previously given, an expression may be obtained for the field intensity in the gap, that is:[3]

$$B_g = 4\pi J\left(1 - \cos a + \sin^2 a \cos a \log_e \frac{D}{d}\right) \qquad (19)$$

[3] See Weiss, former citation on this matter.

FIGURE 38—1,000-Kilowatt Arc Converter (Anode Side)

Figure 39—1,000-Kilowatt Arc Converter (Cathode Side)

which of course is a maximum when $a = 54^\circ 44'$ for cones the apices of which coincide. In this expression, $1 - \cos a$ represents the flux due to the pole tip face, and the remainder that due to the conical surface.

Since the flux distribution in a pole tip of any shape varies with the flux density and with the permeability of the steel, any mathematical treatment of the problem which seeks to render the computation of B_g possible thruout the broad range of flux densities from 0 to 40 or 50 kilogausses, must be based upon these premises, which are incapable of exact mathematical expression. Therefore a mathematical solution of the problem for this broad range is seemingly impossible.

The range of flux densities used in arc converters is from approximately 2 to 20 kilogausses, and hence we must resort to experimentation upon actual magnetic circuits to obtain arc design data.

BEST PRACTICAL TIP-GAP RATIO $\frac{d}{G}$

The mathematical derivation of the pole tip angle $a = 54° 44'$, Figure 35, makes the ratio of pole tip face diameter "d" to gap "G" equal to $\sqrt{2}$ for cones with the same apex, since $\tan^{-1} a = \sqrt{2}$ and $\tan a = \frac{d}{G}$.

Since commercial arcs cannot have the mmf. applied ideally, that is, the arc chamber cannot be within the magnet windings, there is more spreading of the flux in the gap than in the ideal case. Hence $\frac{d}{G}$ must be greater than the theoretical value, namely, $\sqrt{2}$. The practical value seems to be very close to $\tan 60° = \sqrt{3}$. However, the $\frac{d}{G}$ ratio may be varied between 1.6 and 2.0 without seriously reducing the flux density in the gap. Figure 40 gives experimental proof of the foregoing.

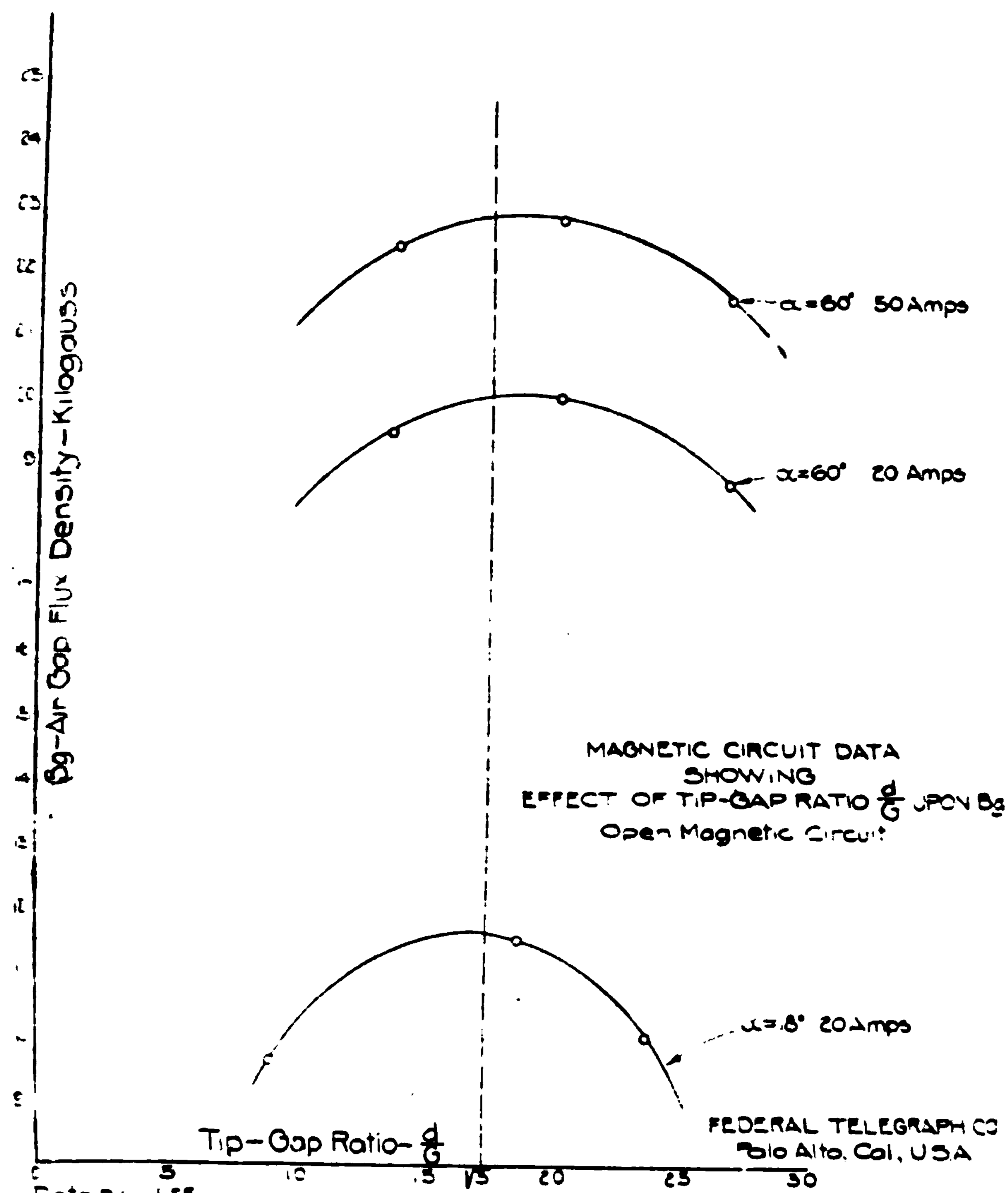


FIGURE 40

In all the work described herein, magnetic flux densities were measured by means of an exploring coil and ballistic galvanometer, in the usual well-known manner.

Best Practical Pole Tip Shape

The fact that the *best practical* ratio $\frac{d}{G}$ is very close to $\sqrt{3}$ shows that the angle of the two cones, the tips of which touch in the center of the air gap is 60°. However, as shown on Figure 41 and in Table 3, 60° tips are not as good as 55° (keeping $\frac{d}{G} = \sqrt{3}$ in both cases).

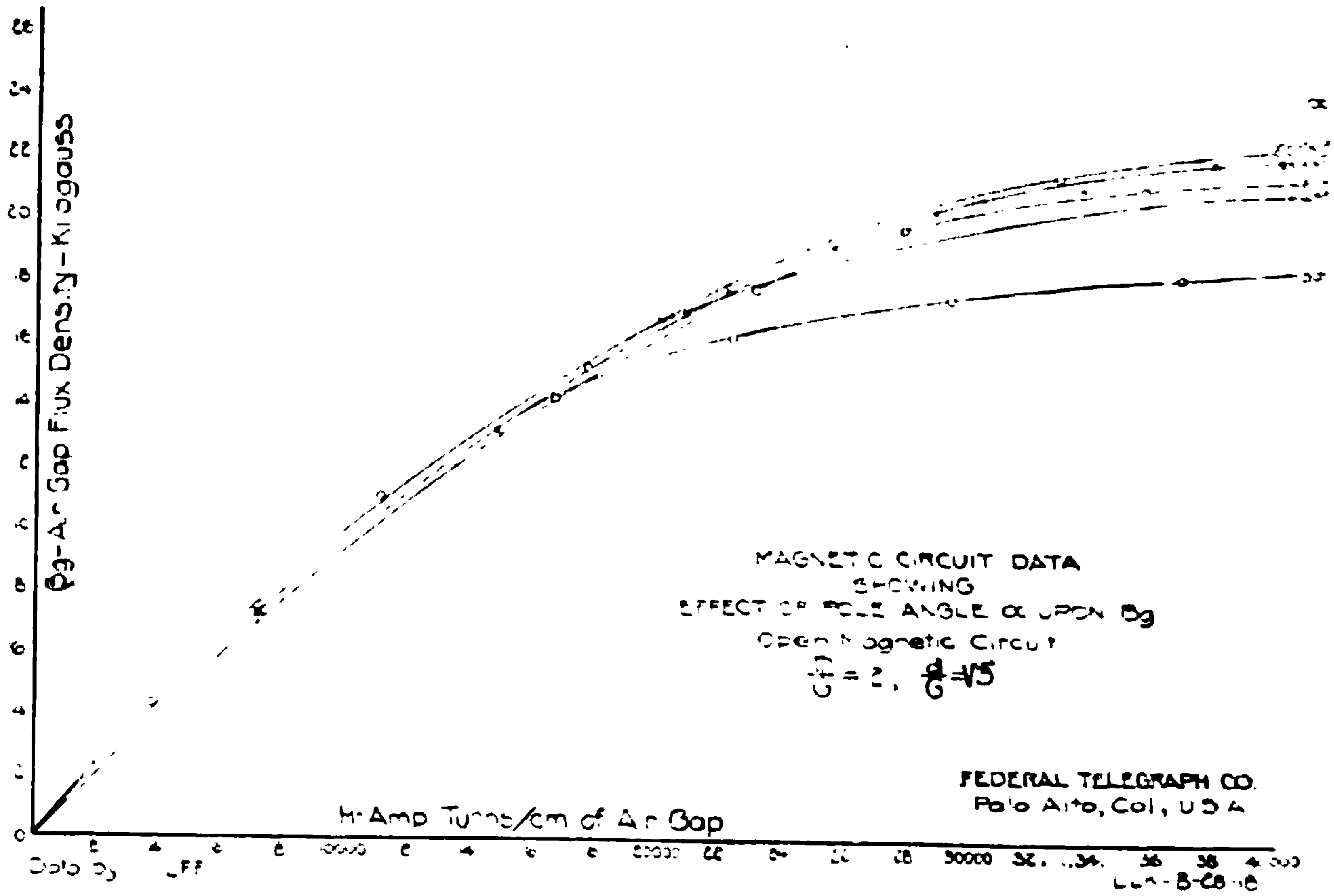


Figure 41

TABLE 3

(Figure 41—Pole Tip Shapes)

Open Magnetic Circuits

H	B	
3,850	4,320	
7,260	7,440	
11,200	11,050	$\frac{D}{G}=12$
17,800	15,290	
18,010	15,380	$\frac{d}{G}=\sqrt{3}$
21,780	17,890	
25,320	19,500	$a=60°, 55°, 50°$
32,800	21,400	
39,800	22,300	
43,490	22,700	
7,230	7,330	
14,840	13,250	
20,000	16,900	
20,900	17,080	$\frac{D}{G}=12$
22,250	17,880	
25,600	19,220	$\frac{d}{G}=\sqrt{3}$
28,900	20,300	
30,300	20,800	$a=55°$
33,500	21,100	
37,800	21,850	
39,500	21,900	
7,350	7,030	$\frac{D}{G}=12$
16,700	14,280	
23,100	17,850	$\frac{d}{G}=\sqrt{3}$
27,800	19,800	
35,600	21,000	$a=60°$
7,500	7,620	
14,770	13,050	$\frac{D}{G}=12$
22,400	16,280	
29,400	17,500	
36,800	18,100	$\frac{d}{G}=\sqrt{3}$
44,500	18,600	
60,700	19,150	$a=30°$
7,500	7,850	
14,700	13,260	$\frac{D}{G}=12$
22,300	17,750	
29,580	19,620	
30,850	19,890	$\frac{d}{G}=\sqrt{3}$
37,450	20,700	
38,050	20,750	$a=45°$

The foregoing indicates that the best *practical* tips should start at 60° near the gap and change to approximately 55° part way back on the cone.

In view of our previous discussion of the Weiss tip of Figure 37, the 55° angle should be reduced near the base of the cone Such a tip is 60°—55°—50° (approximately one-third of the slant height for each angle).

The experimental results of Figure 41 prove the correctness of this reasoning, since the 60°−55°−50° curve lies above all others.

It is probable that some modification of these angles would give a somewhat better tip, but itmust be remembered that it is inadvisable to have a greater portion of the sloping pole surface below 55° than in the tip just considered, because at the rate at which it increases tip height and hence arc weight and dimensions. These reasons practically prohibit the use of any angle less than 50°, and in view of these practical considerations and the ease with which the 60°−55°−50° tip may be machined, it is considered one of the best all-around designs for high power arc work.

Best Practical Pole-Gap Ratio $\frac{D}{G}$

According to equation 19, $B_g \propto log_e \frac{D}{d}$, all other factors remaining constant. Since $d = \sqrt{3}G$ for best results, we would expect $B_g \propto log_e \frac{D}{G}$. Figure 42 shows a curve plotted to this equation.

Obviously this ratio is most important, for it is the main factor controlling the weight and cost of an arc converter. If it is too low, the magnetizing ampere turns are not working to full advantage; and if it is too high, both steel and copper are wasted because of the excessive mean diameter of the magnet winding.

Of course, the most economical weights of steel and copper are dependent upon the market price of each. In general, however, *the magnetic circuit of an arc converter should be designed so that the knee of the saturation curve is reached at rated full load current.*

In accordance with this principle saturation curves have been taken over a great range of $\frac{D}{G}$ ratios for various pole tip angles.

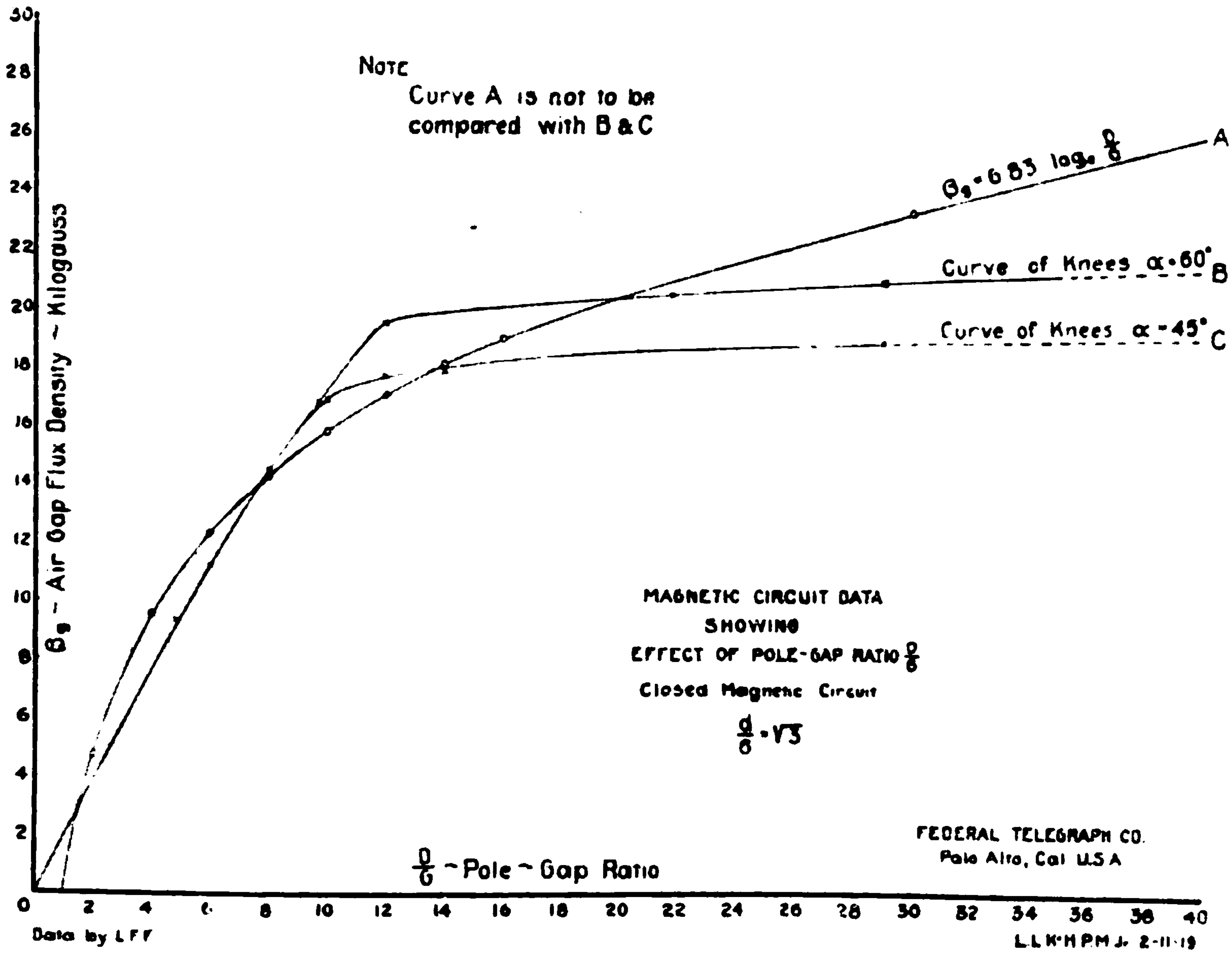


FIGURE 42

The knee of each of these curves has been determined, and B_g at the knee tabulated with $\frac{D}{G}$ and a. In all cases $\frac{d}{G}$ was held constant at its best value $\sqrt{3}$.

Table 4 gives typical data of this sort, which are shown graphically on Figure 42 for $a = 60°$ and $45°$. The difference in the shape of these curves and the $log_e \frac{D}{G}$ curve obtained by computation is due to the fact that the ordinates of the experimental curves are B_g *at the knee* which introduces a *variable H* into their past history; whereas the computed curve is for an arbitrarily chosen *fixed value of H*. Since we are interested primarily in the *position of the knee* of the saturation curve, the computed curve is of no particular value in design.

TABLE 4

(Figure 42)

$\frac{D}{G}$	B_g (Knee)		
4.83	9,300		*Observed Data* on Effect of $\frac{D}{G}$ on *Position of Knee* of Saturation Curves with *Variable magnetizing force*
6.0	11,160		
9.75	16,730	$a = 60°$	
12.	19,500	$\frac{d}{G} = \sqrt{3}$	
21.8	20,500		
29.	20,900		
35.	21,000		
6.	11,160		
8.	14,410		
10.	16,820	$a = 45°$	
12.	17,660	$\frac{d}{G} = \sqrt{3}$	
14.	17,850		
29.	18,850		

$\frac{D}{G}$	$B_g = 6.83 \log_e \frac{D}{G}$	
2	4.74	*Computed Data* on Effect of $\frac{D}{G}$ on B_g at a *fixed arbitrarily chosen magnetizing force*
4	9.50	$a = \text{constant}$
6	12.22	$\frac{d}{G} = \text{constant}$
8	14.2	
10	15.75	
12	17.0	
14	18.05	
16	18.95	
30	23.25	

Two important conclusions may be drawn from these data.

(1) When $\frac{D}{G}$ is small, the tip angle a is of little importance. Thus in small arcs where weight is often a controlling factor and $\frac{D}{G}$ is therefore made low, any a that works in well with the chamber design may be used.

(2) $\frac{D}{G}$ should not be made greater than 12. As a matter of fact, considerations of cost and weight make it advisable to run

$\frac{D}{G}$ below its best value in high power arcs. For example, the 1,000-kw. arc converter requires 15,000 gauss (under usual specifications) in a seven-inch (17.8 cm.) magnetic air gap. Figure 42 shows $\frac{D}{G}=8.5$ for 15,000 gauss at the knee of the saturation curve. This makes $D=59.5$ inches (151.1 cm.), whereas the arc as designed has $D=$ 45 inches (114.3 cm.), which saves many tons of steel (counting yoke and bedplate), without increasing the amount of copper, because of the reduction in the mean diameter of the winding.

In all the foregoing, $\frac{D}{G}$ has been considered as constant thruout the entire length of the magnetic circuit. Experiments show that it cannot be reduced appreciably in any part of the bedplate or pole pieces, but it may be reduced somewhat in the yoke, because leakage reduces the total flux which must pass thru the yoke. The cross section of the yoke of the 1,000-kw. arc is equivalent to a pole diameter of 40 inches (101.6 cm.), whereas $D=45$ inches (114.3 cm.) in the main pole and bedplate.

In general, B in the lower portion of the pole or bedplate equals B_g approximately, while B in the yoke or at the base of the pole tip equals approximately $\frac{1}{2}$ B_g.

Open Compared with Closed Magnetic Circuits

It is not customary to have cranes at the point of installation of arcs up to 100 kw., and the weight of the unit is a matter of considerable importance. This is especially true on shipboard.

Unless a closed magnetic circuit has a fairly large $\frac{D}{G}$ ratio, and is therefore heavy, it is of little value. Open magnetic circuits are used exclusively in sizes up to and including 100 kw. because a pound of copper is worth several pounds of steel in producing B_g, and the magnetizing kilowatts are small in any case. Furthermore, in these small sizes it is often found that at the present market prices of steel and copper, a saving in cost is effected by using the open magnetic circuit.

The effect of the arc on a ship's compass is a factor which sometimes makes it desirable to use a thoroly shielded closed circuit design when otherwise the open circuit would be preferable.

It is not necessary to build a full size magnetic circuit to determine the shape of the saturation curve. The use of models is a quick and inexpensive means of accomplishing this result.

Figure 43 and Table 5 show how well this can be done. It will be noted that the ratio of the weight of the full-sized magnetic circuit to its model was 650-to-1, and that the measured flux densities of the two lie so close together that it has been possible to draw but one curve thru the two sets of points.

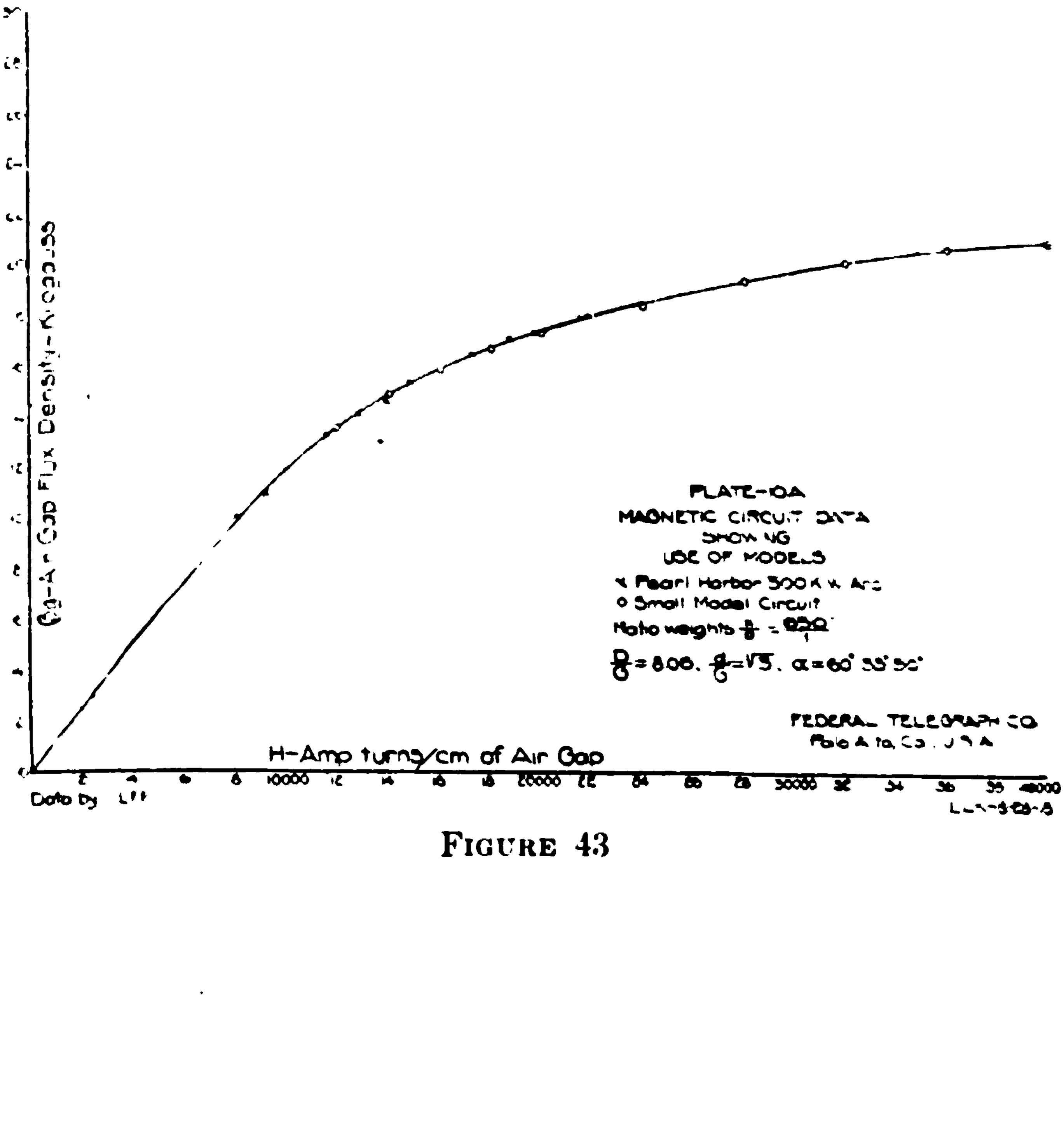


FIGURE 43

TABLE 5

Figure 43—Model Data

$$\frac{D}{G}=8.06;\quad \frac{d}{G}=\sqrt{3};\quad \alpha=60°-55°-50°$$

Pearl Harbor 500 kw. Arc.		Magnetic Model	
H	B_g	H	B_g
2,500	3,100	12,000	13,700
3,800	4,650	14,000	15,700
6,700	8,200	16,000	16,100
8,200	10,100	18,000	16,900
9,300	11,000	20,000	17,500
11,600	13,400	22,000	18,100
12,800	14,300	24,000	18,500
13,900	14,800	28,000	19,500
14,800	15,600	32,000	20,200
16,100	16,100	36,000	20,700
17,200	16,700	40,000	20,900
18,700	17,200		
19,000	17,300		
19,700	17,500		
20,700	17,800		
21,800	18,100		
23,200	18,600		
Ratio of Weights 650:1			

Figures 44, 45, and 46 show these magnetic circuits. It will be noted that the mechanical form of the two circuits is as radically different as their size, and yet they are excellent magnetic equivalents.

FIGURE 44—500-Kilowatt Arc Converter (Anode Side)

FIGURE 45—500-Kilowatt Magnetic Circuit

FIGURE 46—Model Magnetic Circuit

DETERMINATION OF THE MAGNETIC AIR GAP *G*

The determination of the magnetic air gap is one of the most important decisions necessary in the design of a large arc converter. It is dependent upon oscillatory circuit resistance, kilowatts input, and, in fact, upon nearly all the specifications which must be given the engineer before design work may be started. The shape of the anode tip is also an important factor.

Analytical attack on this problem has not been possible thus far. All progress has been made by experimentation. It is believed that an analysis of the results would be too cumbersome to include in this paper.

In general it may be stated that the air gaps now in use range from 1 to 7 inches (2.54 to 17.78 cm.).

Determination of Chamber Size

Figure 47 shows the area of chamber cooling surface which is considered good practice at this time. All arcs of modern design are constructed in accordance therewith.

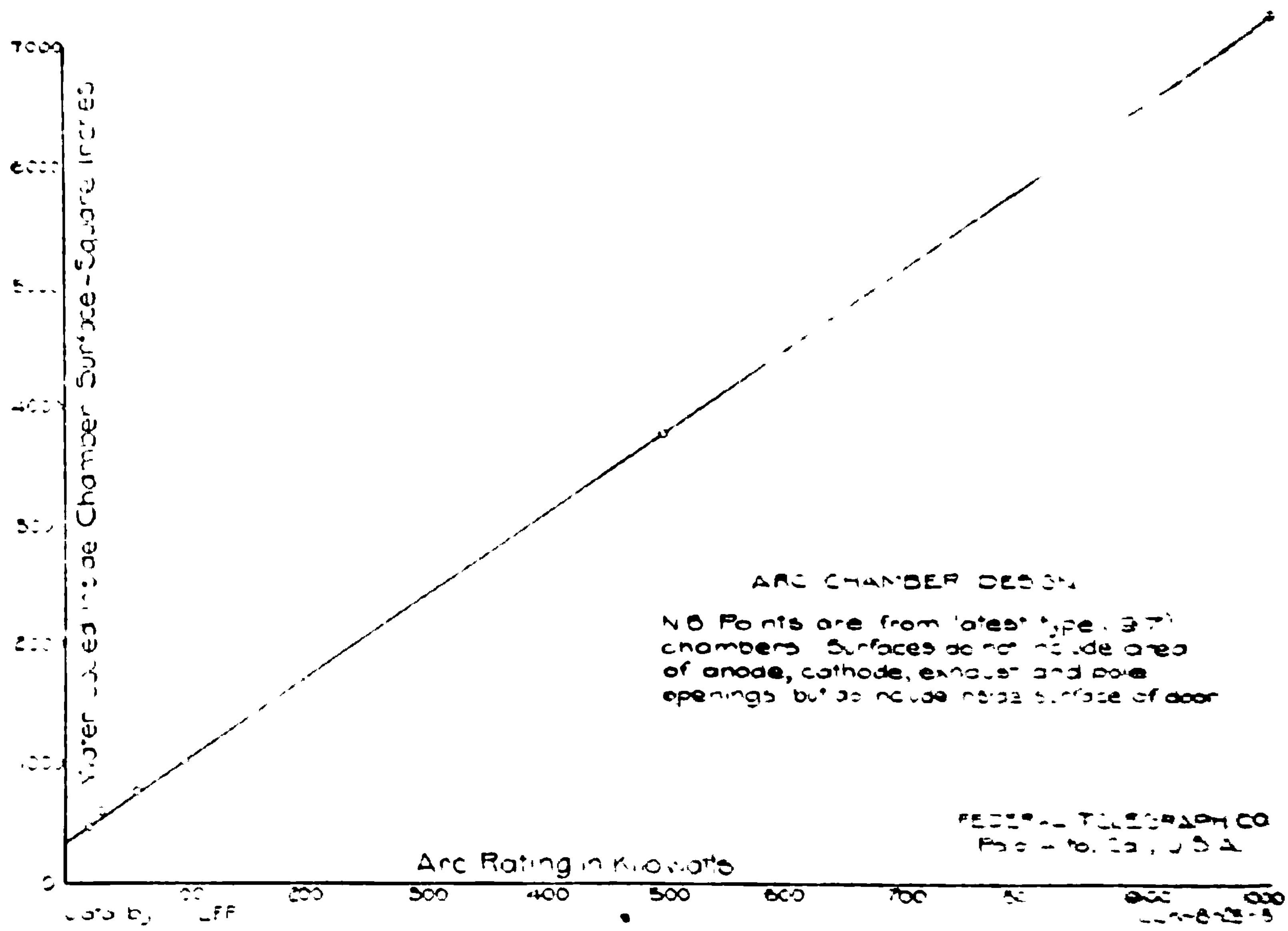


FIGURE 47

Once the necessary area has been determined, the design of the remainder of the chamber is one of purely mechanical problems.

The size of the water jackets is determined by the difficulties of the foundryman to a very considerable extent, as the casting must be watertight.

The thickness of water jacket which is satisfactory to him in his core work is usually of ample cross section to permit the water flow necessary at the usual 15-20 pounds per square inch (1.06 − 1.41 kg. per sq. cm.) pressure available.

Choice of Arc Electrodes

In the ordinary carbon arc, particles are carried over from the positive electrode and deposited on the negative. This causes the formation of the well-known positive crater and negative cone.

Some arcs however (such as the magnetite arc used for illumination), use copper anodes, which do not wear away at an appreciable rate. The Poulsen arc is of this type.

Carbon or graphite may be substituted for the copper anode with excellent results, but these materials have never come into regular use because a water-cooled copper electrode is less troublesome.

They are the usual cathode materials, however. Carbon should not be operated above 200 d.c. amperes per square inch (31 amperes per sq. cm.), and graphite twice this density at the usual wave lengths encountered in present day radio telegraphy.

The d.c. amperes may be used as the basis for computing the cathode diameter safely, because the skin effect in materials of high specific resistance is relatively small. For example, at a wave length of 3,000 meters a carbon electrode 1.60 cm. (0.63 inch) in diameter[4] has a radio frequency resistance but 1 per cent. greater than its d.c. resistance. At a wave length of 9,000 meters, this diameter may be increased to 2.77 cm. (1.1 inch). Thus the determination of electrode diameter on the basis of a limiting direct current density does not vary appreciably from the diameter which might be computed with the radio frequency current as a base.

Acknowledgments

In conclusion, I wish to express my appreciation of the assistance of Mr. R. R. Beal for taking flux data, of Mr. A. Anderson, for tabulating the same, and of Mr. W. A. Hillebrand for many helpful criticisms and suggestions.

SUMMARY: The Poulsen arc converter cycle is studied in detail, and the various relations between direct and alternating arc currents and voltages obtained and discussed. The possible production of harmonics is considered.

The effect of arc field strength on the arc phenomena is then taken up. The efficiency of the Poulsen arc cycle is obtained and checked experimentally.

The theoretical relation between best field strength and wave length, arc current and voltage, and nature of atmosphere surrounding the arc, is deduced, and shown to be correct by elaborate experimental data. In this connection, ethyl alcohol and kerosene atmospheres are compared.

The design of the magnetic circuit is then handled in detail. The most economical pole shape is given, the tip being a triple tapered conical frustrum. The tip-gap ratio and the pole-gap ratio are also experimentally obtained.

Open and closed magnetic circuits are compared. The design of these large electromagnets is facilitated by the use of small models, the results thus obtained being reliable. Data are given for desirable chamber surface for various arc inputs, and for the design of arc electrodes.

[4] Circular 74, 1918, Bur. Stds., pages 299-310.

THE UNI-CONTROL RECEIVER*

By

Roy E. Thompson

(Chief Engineer and General Manager, Wireless Improvement Company, New York)

The original paper on the "Uni-Control Receiver" was prepared some time ago, and arrangements were made to read it before The Institute of Radio Engineers shortly thereafter. Certain delays interfered until the United States entered the war, so that it was no longer considered proper to read it.

In order to show some of the causes which led up to the design of the receiver originally, I wish to point out that—

The law says in effect to people who instal and operate radio apparatus: "You must provide only apparatus which is capable of selective operation; that is to say, your apparatus must be sharply tuned so that if you are interfering with another station he may tune you out by slightly changing his receiving tune." Then the same law goes on and says in effect: "In view of the fact that all transmitters have now been required by law to be sharply —that is selectively tuned—it is feared that in case of distress a vessel may call for help and not be heard. So, to eliminate this possibility, it is decreed that you must handle your business on one of two wave lengths, namely, 300 or 600 meters. That is, we make you tune sharply so that listeners on slightly different tunes will not hear you; then we say that you shall all send on the same tune so that all listeners *will hear* you."

This provides a situation which in practice is no doubt different from anything the framers of the laws anticipated when they agreed upon the London Convention in 1912.

Now there was a very good and sufficient reason as to why these two major requirements of the law should have been enacted. The flaw in the scheme is that the carrying out of the one negatives the results expected from the other and vice versa.

It was with the above conditions that I and other radio inspectors of the Department of Commerce tried to cope when the

* Received by the Editor, January 23, 1919. Presented before the Institute, New York, February 5, 1919.

laws became effective on December 13, 1912. It has been said that one never knows the weakness of a law until one tries to enforce it. In this case, we did not fully realize the weakness of the law until after it was fully enforced. We first berated the framers of the law, but upon sober thought we were compelled to acknowledge that it seemed practically impossible to frame a law which would effectively regulate radio communication and secure the desired results, or for any company or government to formulate rules which would properly take care of the conflicting interests represented on one side by the commercial business that must be handled and on the other by the desire and necessity of keeping this commercial business from interfering with possible messages relating to the safety of life at sea.

The thing back of this whole problem—the thing that caused the attempt to solve the problen by regulation and the thing that in turn prevented regulation from being the solution—was plainly the limitation of the radio apparatus itself. The transmitter had to be restricted as to the number of wave lengths it could use so that the receiver might always hear it, and the receiver had its selective ability nullified by being compelled always to keep itself adjusted so that it would hear the transmitter on its restricted wave length.

So far as I know, no laws or regulations have ever been proposed or worked out which would solve the conflicting problems referred to above. In my opinion, neither government nor private monopoly would do it because the only advantage that monopoly has over regulation is the power to regulate arbitrarily instead of equitably.

These and the less important problems relating to the human element or the tendency of operators to neglect fully to carry out regulations, instructions or even arbitrary commands were the causes for the writer's determination to design if possible a receiver which would take care of both the conflicting requirements of the radio laws and permit the emancipation of the transmitter by producing a receiver which would hear everything within a space of a few seconds, or before much harm could come to a distressed vessel and still be able to exclude anything which the present day receivers will exclude.

The first requisite was a receiver that could be caused to "sweep the ether," so to speak, in the same way as a searchlight sweeps the horizon. This means that there must be a gradual and continuous sweep from the shortest to the longest wave length within the range of the instrument. A receiver

embodying this faculty in a very simple form is one having only one variable point or element. A fixed inductance and a variable condenser in series with the antenna and ground and with the detector and telephones across the inductance or condenser is a simple illustration. Such a device has not sufficient range to fill the practical requirements however. As a practical matter, it develops that more than one varying or wave changing point must be provided.

The second requisite was that in whatever direction the receiver developed, it must be operated entirely by a single driving force or control. It must further be capable of repeating the wave change cycle over and over again at a speed permitting the operator's attention to be attracted by faint signals whenever the receiver swept by or passed thru the tune corresponding to the wave length of such signals.

Such a device, I believe, is the receiver to be described in this paper.

An elementary diagram illustrating the principle is shown in Figure 1, which shows a unit coil of inductance UL and a multiple unit coil of inductance ML. A capacity K provides for the oscillatory character of the circuit. The unit switch US has attached to its shaft a gear such as the spur or intermittent type, which drives the multiple unit switch MS thru a similar gear of slower rotation. As the unit switch brings consecutively into the circuit the inductance represented by taps *a*, *b*, and *c*, respectively, the multiple unit switch moves over to tap 1, 2, or 0 as the case may be. The movement of the multiple unit switch from one tap or segment to another is arranged to take place whenever the unit switch moves from tap *c* to *d*. The unit switch

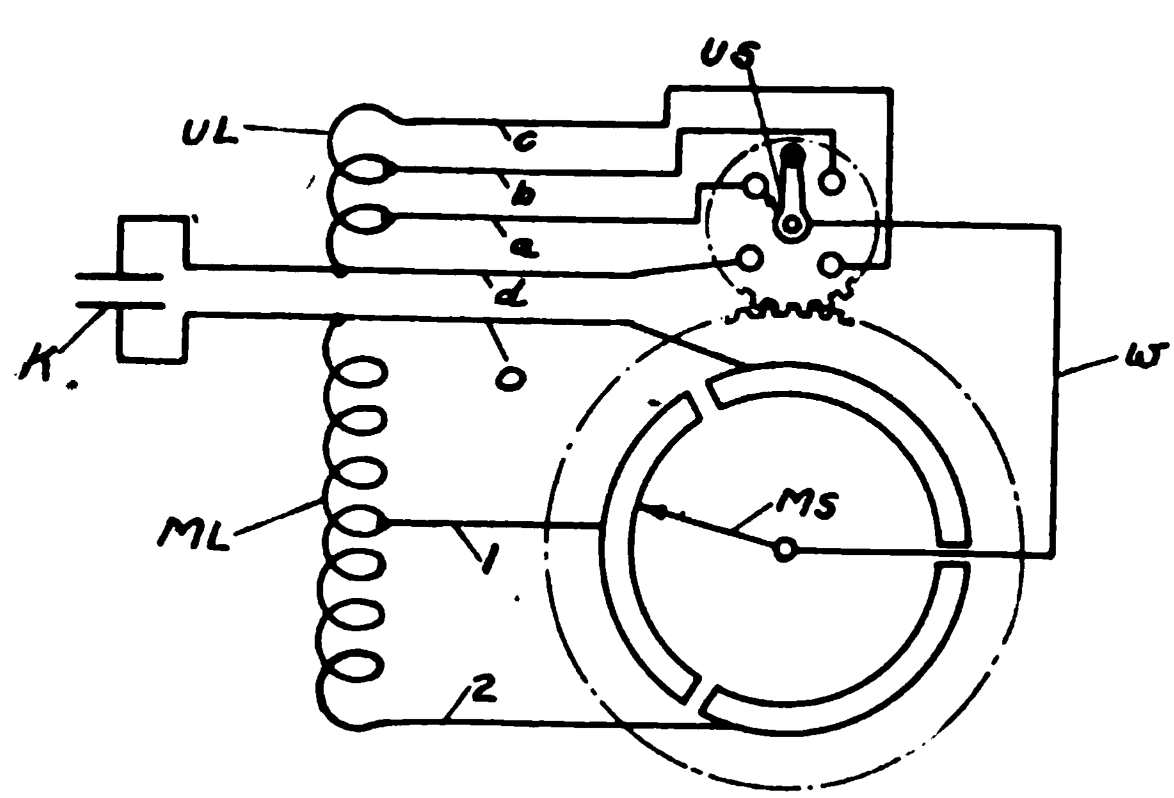


FIGURE 1

moves clockwise and the multiple unit switch counter clockwise.

Figure 2 illustrates the same circuit as Figure 1 with the addition of a variable condenser VC which enters the circuit in series with the inductance coils whenever the switch arm MS connects the tap or segments 0 and 3.

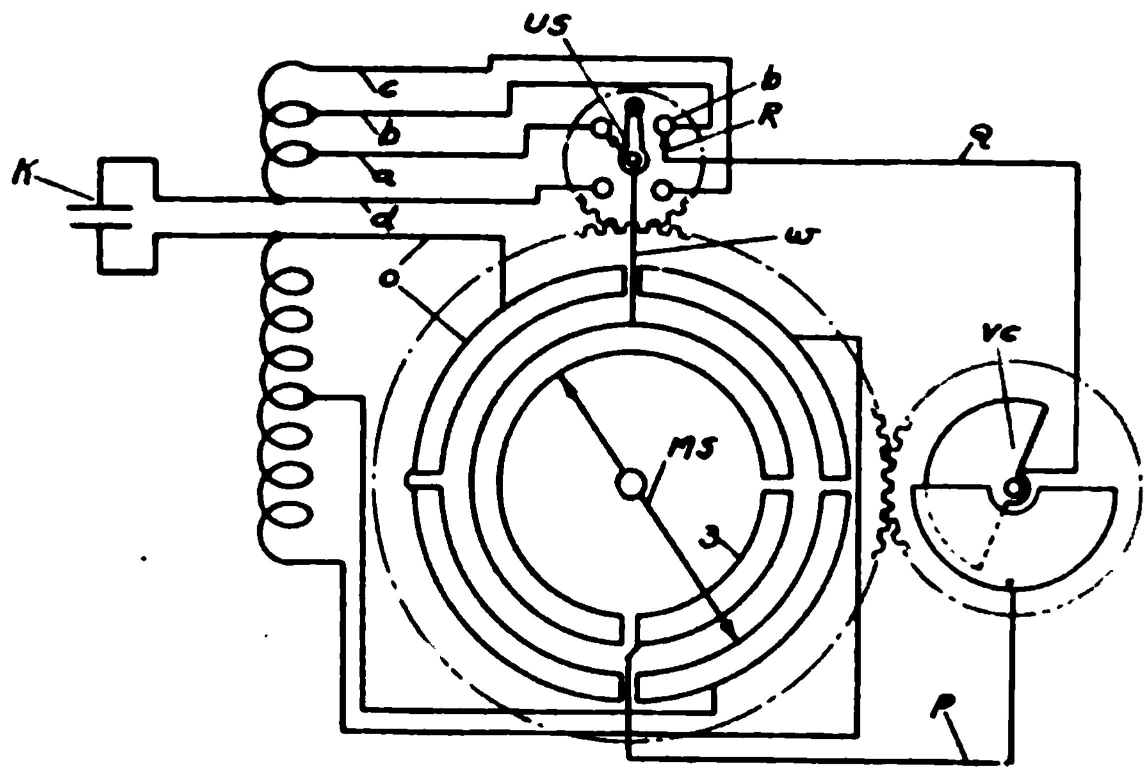


FIGURE 2

It will be noted that whenever condenser VC is in circuit, the circuit instead of being thru the return wire W and the tap a, is thru the wires P, Q, and the adjustable contact R, and the tap b. As the contact R can be placed on any of the taps a, b, c or d, it can be placed so as to compensate for the entrance of the condenser VC into the circuit and keep a wave length gap from appearing there. Of course, in practice there are always a few turns of inductance kept in the circuit for coupling to the detector. With the compensating inductance controlled by contact R the receiver may be adjusted for use with a condenser K having different capacities. As the condenser K represents the capacity of an antenna, the set may be adjusted for different antennas so as to have no wave length gap or overlap when the variable condenser VC enters the circuit for the purpose of reducing the period of the circuit below its normal or fundamental wave length, such as when it is desired to come down to short wave lengths.

Figure 3† is the same as Figure 1, with the condenser of Figure 1 replaced by the antenna F and the ground G, and there is also

†The method shown in Figures 3 and 4 for securing oscillations of the period of the oscillatory circuit is only one means of securing the desired result. Any other or more preferred method for exciting the antenna or oscillatory circuit at substantially its own period may be used.

connected to the inductance a generator of radio frequency currents in such a manner as to generate waves of substantially the same length as the wave length of the antenna. The signaling device *S* may be a telegraph key or telephone transmitter. This figure illustrates the device as applied to a transmitter.

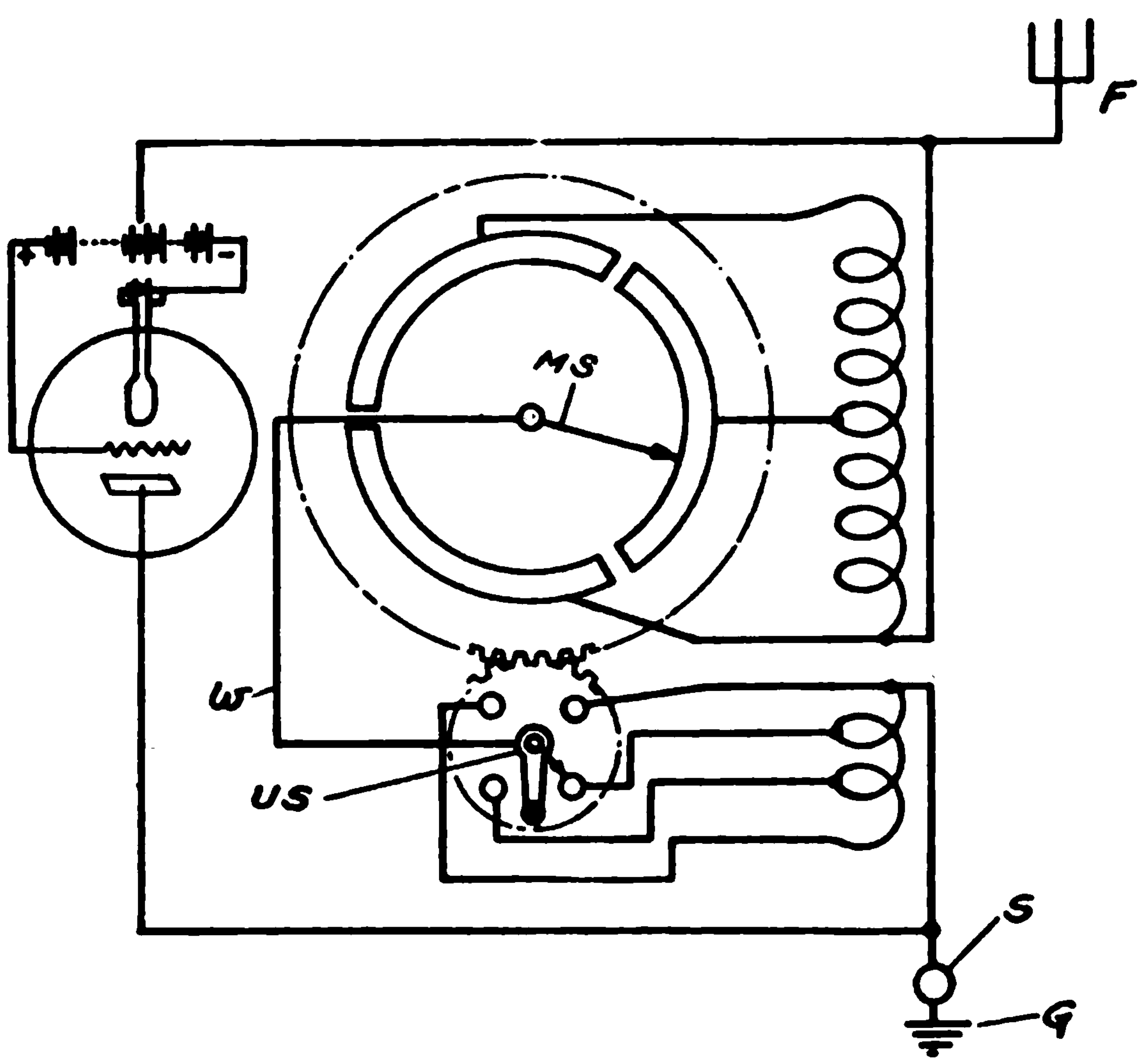


FIGURE 3

Figure 4† illustrates the same circuit as Figure 3, with the addition of a detector circuit for indicating such damped wave trains as may be produced in the antenna *F-G* by the arrival of damped wave trains or continuous wave signals, either or both, from a distant station.

Figure 5 represents a wiring diagram of a receiver having a range from 150 up to about 3,500 meters on the ordinary ship's antenna.

The fixed or minimum number of turns in the antenna coil is 10 turns. The total turns controlled by the unit switch is 25 turns and therefore each section of the multiple unit coil has 26 turns. Every other section is cut off from the rest by a dead-end cut-off switch.

The ratio of the gears of the unit switch to the multiple unit switch is 1-to-12, and between the multiple unit switch and the

† The method shown in Figures 3 and 4 for securing oscillations of the period of the oscillatory circuit is only one means of securing the desired result. Any other or more preferred method for exciting the antenna or oscillatory circuit at substantially its own period may be used.

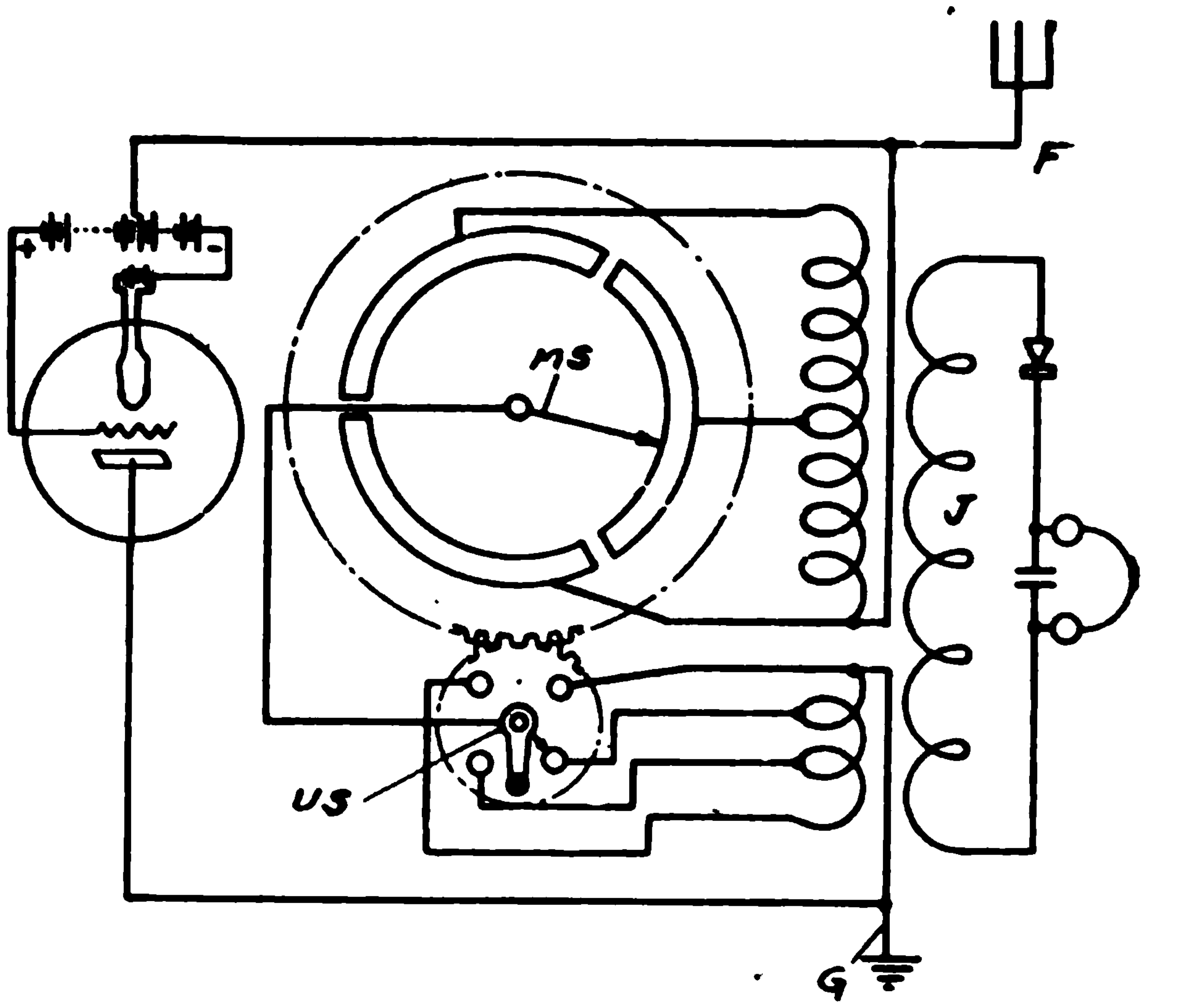


FIGURE 4

variable condenser is 3-to-1. That is, the multiple unit switch travels one-sixth of its circumference on the segments 3 and 0 while the variable condenser is going from 0 to 180 degrees.

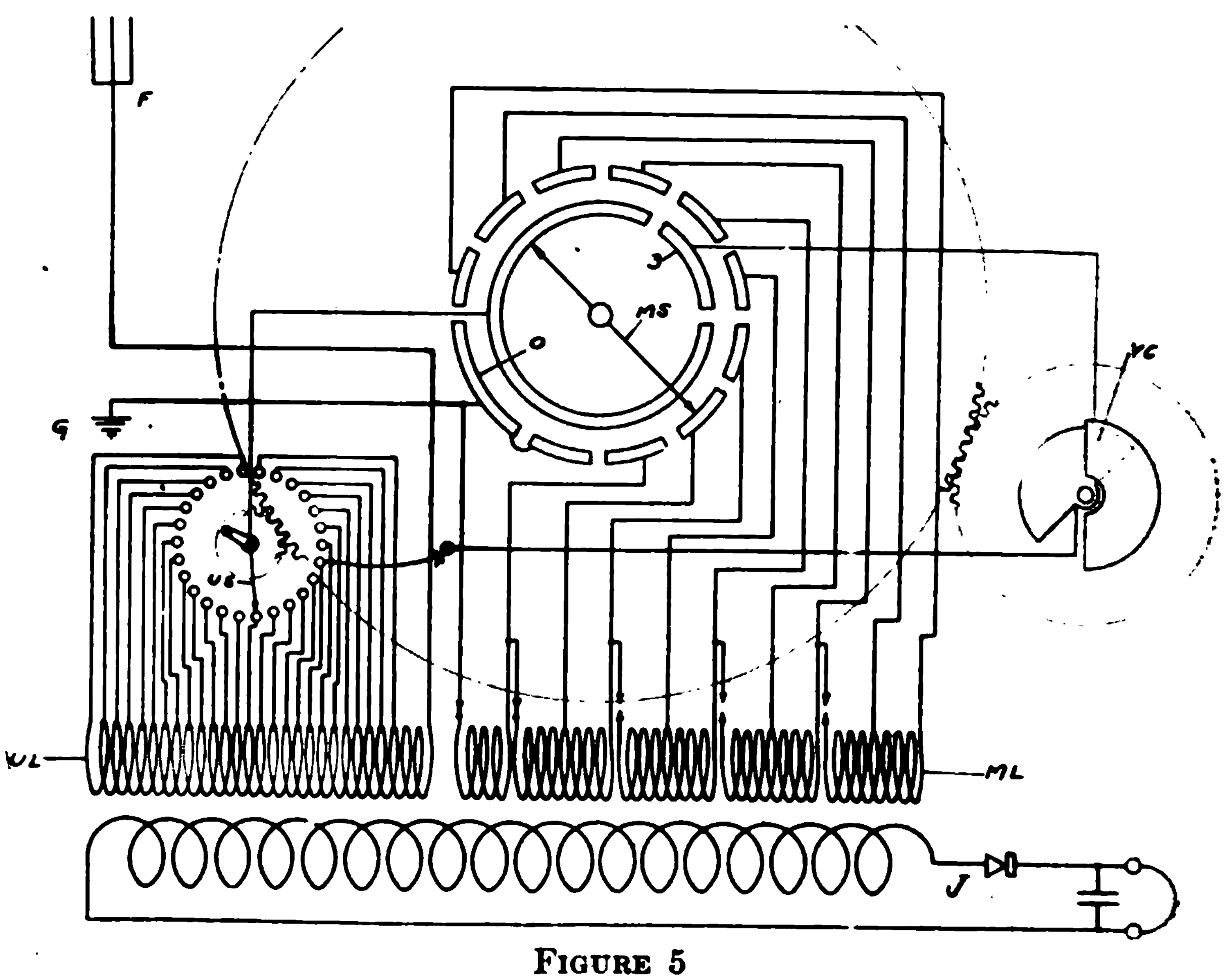


FIGURE 5

Tho the condenser keeps revolving while the unit switch is varying the wave length, it is not in circuit except during the above described period. Also, tho the unit switch keeps revolving while the variable condenser is in circuit, the circuit is then no longer thru the unit switch so that the inductance is not varied.

If the detector circuit *J* is properly constructed and associated with the antenna inductance coils, the antenna is left free to oscillate without interference or reaction by associated circuits and the full advantage may be taken of antenna resonance. The detector coil and circuit as a whole must be substantially non-accumulative. That is to say, the capacity which gives the ordinary coil and associated elements the power to accumulate energy in substantial quantities to react against the antenna or exciting frequency must be eliminated to a point where its effects are no longer felt and then if this coil and the antenna coils are associated with the proper degree of intimacy, the antenna and detector coil act entirely as one at all frequencies and the maximum combined efficiency and selectivity are obtained without the many objectionable features of attaching the detector physically to any part of the antenna circuit or attaching it to an associated tuned circuit.

I have made the essential requirements for the construction and association of such coils and circuits so as to obtain the maximum efficiency and selectivity the subject of separate patent applications, and will be glad at some future date to read a paper and describe in detail the principle and construction involved whereby I am able to secure for any wave length range equal or greater selectivity and efficiency than is secured by the most modern two-circuit receivers.

Figure 6 is a photograph of a type of receiver furnished to the Signal Corps and is intended as the ordinary commercial or ship type. It is to be noted that the detector coil is movable with respect to the antenna coil. This, however, merely has the effect of decreasing or strengthening the signals, useful in congested areas where one does not always want the fullest strength of signals obtainable. A hand crank only was provided with these particular receivers.

Figure 7 was taken with the front panel of the receiver off and shows the dial calibration card with the wave length space blank. When the receiver is attached to a particular antenna this calibration card is taken off and the wave lengths corresponding to the divisional lines are marked in the space provided. This card is then properly adjusted on the receiver with respect

FIGURE 6

to the positions of the switches and incoming wave lengths are then read directly.

Figure 8 shows the calibration card removed, exposing the gear drive which is self-contained in an aluminum casting frame. The large gear is of bakelite.

FIGURE 7

FIGURE 8

FIGURE 9

Figure 9 is a top view showing antenna coils, variable condenser, dead-end cut-off switches, and so on. All apparatus is mounted on the middle or bottom panels.

Figure 10 shows the motor drive and worm gear box for driving the receiver thru an especially designed flexible speedometer drive. A speed control rheostat and start and stop push button is also shown on the table.

FIGURE 10

The schematic arrangement is shown in Figure 11. The apparatus to the left of the thin dotted line may be under the table or some place out of the way of the operator. The apparatus to the right of the dotted line should be in easy reach of the operator. The clutch mechanism whereby the operator engages or disengages the power drive is best shown in this figure. The flexible power drive shaft is connected to the shaft 10 which thru the bevel gear 11 drives the bevel gear 12. This gear 12 rides free on the main or hand driving shaft 15 except when the teeth 13 are engaged with similar teeth 13A on the collar 15A which turns with the shaft 15 but which can be moved in or out by the operator to engage or disengage the power drive as desired.

In December, 1918, Commander Hooper, while testifying before the Congressional Committee on Merchant Marine and Fisheries which was hearing the Navy's evidence as to why it thought that it should have a monopoly of all radio stations with reference to ship to shore business said:

> "It was found, after the stations became sufficiently numerous, that when one station was working another station would be interfered with, not because there were not a large number of wave lengths in this area of wave lengths, but in order to guarantee that all ships can intercommunicate and communicate with any shore stations they happen to pass in any part of the globe, they must all communicate on one wave length, which happens to be designated as 600 meters. That wave length is agreed to internationally and is in common use thruout the world. If that had not been agreed upon, it is obvious that ships choosing any wave length they wished to call the shore station, and the shore station listening on all wave lengths at the same time, they would not hear the calls. A ship in distress might require hours before she could reach anybody, and it would be a very serious situation. In fact, it is absolutely necessary that all ships and all coastal stations which are able to work with the ships should call and answer each other on 600 meters. Now, since that is necessary, it must be a matter of very careful regulation to see that they do not interfere with one another. For example, a ship on the high seas, 300 or 400 miles (500 to 700 km.) off the coast, wishes to send a message to Norfolk at the same time that Norfolk wishes to send a message to the ship. He calls Norfolk, and Norfolk tries to send a message to him on 600 meters, and Charleston begins to send a message to some other ship on 600 meters, and this poor fellow is out there where the signals come in equally strong from both stations, and he cannot read either. With hundreds or thousands of ships along the coast and many coastal stations all trying to work on the same wave length, it can readily be seen that there is endless confusion unless the most careful regulation is exercised."

Mr. Bankhead: "What sort of regulation do you suggest, in that connection?"

Unfortunately Commander Hooper apparently did not understand the question and his answer therefore did not give us the solution. The laws of selectivity as at present understood do not permit a multiplicity of stations to transmit on the same wave length without interference regardless of how sharp the transmitted wave or how selective the receiver.

Commander Hooper, still speaking of this phase, said:

"If it were not for the fact that all the ships have to listen for calls on the same wave lengths and that the apparatus should not be tuned too sharply—otherwise the man listening will not hear the ship call, even if he has nearly the same wave length—then there would be no question in it as to this second phase. But the ships must listen on this 600 meters, and the shore station must listen on that, otherwise their calls will be unheeded. I think it is obvious that to get the best service we must have it under the most highly organized control. I do not think there will be any objection to this on the part of the commercial companies. Their objection is going to be on the third phase."

It is plain from Commander Hooper's statements that the most convincing argument for Government ownership is this impossible situation of all ship-to-shore or ship-to-ship business having to be carried on and all listening and all calling being done on 600 meters. This is true even tho the proponents of Government ownership do not clearly explain, in the light of the international treaties and regulations and the fact that a great part of the ships causing interference fly foreign flags, just how Government monopoly is going to overcome the difficulty.

It may now sound like a rather far-fetched prediction but I truly believe that if all ships at present had permission to transmit on any and all wave lengths and were in turn all equipped with some such device as the uni-control receiver, if then they all operated in accordance with well conceived international regulations, interference would be reduced far below anything which may possibly be brought about thru government or private monopoly. Not only do I believe this would be true but I further believe that the chances for distress messages going unheard would be likewise reduced, regardless of what wave length the ship in distress chose to send its distress call upon.

Furthermore, it would no longer be necessary for all ships to quit work within the radius of the distressed vessel, except those engaged in helping the distressed vessel. International regulation could simply reserve a certain wave length range for distress business and all other stations not concerned could shift to other ranges and go about their normal business.

Let us see how this might possibly be done.

First, let us assume that the present receiver will do everything claimed for it as regards selectivity and efficiency, or that it can in the hands of proper radioengineers be made to do so. I believe that it is said that with undamped single wave transmission,

stations the wave lengths of which do not differ less than 5 per cent. will not interfere. Now if the transmitters only have a range from 300 to 3,000 meters there are still 48 wave lengths within this range, all of which differ at least 5 per cent. from each other. As I see it, there is no reason then why at least 48 ships or land stations might not send and 48 land or ship stations all receive, or a total of 96 stations work, all within the same radius without using up any more of the total practical wave length range available than from 300 to 3,000 meters. As ships may easily go up to 5,000 or 6,000 meters without getting into the high power station range, it is seen that the theoretical limit has by no means been reached.

Figure 11

The reason why this cannot be done with present apparatus is that unless a station could know on just which one of these 48 wave lengths it was going to be called, it might never hear the station doing the calling. And then if a ship should open up with a distress call it would probably be found that the stations within hearing distance would be listening on almost any and every wave length except the one used by the vessel in distress. If, however, these 48 wave lengths were consecutively and continuously being passed by the ear of listening operators by means of the motor driven uni-control receiver, no one could possibly call without being heard within a few seconds at the latest and once the call was heard and the receiver stopped at that particular wave length none of the other 47 transmitters need interfere.

Further to illustrate the possibilities of eliminating interference thru the use of the uni-control receiver and still permit a large number of stations to work without fear of not being heard in case of distress calls, let us assume that there are within a radius of 500 miles of New York a total of 96 ship and shore stations, all of whom desire to pair off and work at the same time. This means that 48 stations would be sending and 48 receiving after they all got into communication.

Let us take the dial calibration card that goes with each receiver and divide it up into 48 sectors, each sector to correspond to one of the 48 wave lengths with which each transmitting station would be provided. Let us then number these sectors from 1 to 48 consecutively—Sector 1 to correspond to 300 meters and Sector 48 to correspond to 3,000 meters, and the intermediate sectors to correspond to intermediate wave lengths, no two of which would be closer than 5 per cent. from each other. In addition to these sectors, there would still be the division lines on the card and the actual wave lengths in units of 10 or 25 meters would be marked directly on the card.

A ship 300 miles from New York wishes to get into touch with a station on Long Island. He simply listens while his receiver makes a few complete cycles during which he will hear all of the stations sending between 300 and 3,000 meters. He notes the numbers of the sectors in which there are stations working, and he notes the numbers of the sectors in which he hears no signals. For instance, he finds that stations are working on a number of wave lengths, but he notes that when his receiver is passing thru sectors from 16 to 20, inclusive, he hears no sound. He is, therefore, fairly safe in assuming that the wave lengths corresponding to these sectors are not being used within his receiving radius, and he therefore chooses sector number 18 upon which to transmit his call to the Long Island station. When the receiver of the Long Island station passes thru sector number 18 he will hear his call and will hear no other signals provided there are no other stations using this same wave length closer than the ship which is calling him. He would then choose this same wave length corresponding to sector number 18 upon which to answer, or he can choose a sector corresponding to any other wave length which from his observation will cause the least interference in his neighborhood. In any event, a ship station will hear him about as quickly if he choose sector number 44 as if he had chosen number 18 or number 1. The point is that each one chooses a wave length or sector upon which to call

the other which is not being used by anyone within hearing distance. If upon establishing communication, the stations find that they are actually interfering with some other station the choice of one or two unoccupied sectors upon which to try communication again would undoubtedly result in the establishment of communication without any further interference.

All this time, if any station should open up with a distress call, using any wave length of the 48 between 300 and 3,000 meters, any of the receiving stations the receivers of which were still revolving (indicating that they were not working) would hear them by the time their receiver made one complete cycle, and if by chance the transmitters were equipped with uni-control devices as well as the receivers, the distress call could be sent out on all 48 of the transmitting wave lengths within a few seconds to attract the attention not only of those stations which were not working but also those stations which were.

To accomplish these results neither Government nor private monopoly is necessary, but only such international regulations as would compel all ship and general public service shore stations to be provided with the same 48 wave lengths upon which to transmit, and also to be provided with uni-control receivers with dial cards divided into 48 sectors corresponding to the wave lengths of the transmitters.

The main purpose of this paper is to point out some of the results which might become possible thru the introduction of such a receiver as the uni-control into general use.

In order to encourage radio engineers and operating companies to give the matter due thought and at least attempt in some measure to realize some of the results pointed out as possibilities, the Government and such reputable radio concerns as may wish it, will, for a nominal royalty, be given a license under such patents as may issue.

No attempt was made by me to get any department of the Government to purchase or use the receiver. It was spoken of to representatives of the Bureau of Steam Engineering some time ago, but they said they would not be interested in new devices during the war.

The original sample was given to Dr. Austin in order to secure a comparison of this receiver when tested with other receivers available in the Navy Department. His report was that the receiver compared quite favorably with other well constructed receivers as regards selectivity and efficiency but that he could see no particular advantage in the uni-control

feature. This is the first attempt I have made to argue its possible advantages. The only ones sold have been some which were sent to Colonel Krumm in France, who happened to know of my efforts while he was Chief Radio Inspector in the Department of Commerce, and who cabled for them thru General Pershing.

The wave length ranges and the number of wave lengths which might be used without interference, as mentioned herein, are for illustrative purposes only, and are given simply to illustrate the idea and possibilities of this type of receiver in connection therewith.

I have not attempted to go into the advantages of this type of receiver where inexperienced operators only are available, and neither have I attempted to point out the many valuable uses to which a uni-control receiver could be put in simplifying the work of the average ship operator, whether or not it was desirable to use the motor drive feature and take advantage of the possibilities opened up thru its use.

Undoubtedly the mechanical construction of the receiver can and, in the ordinary course of events, would be designed to meet the requirements to which the receiver would be put. I have simply shown circuits and mechanism which will serve all the requirements of a ship or general public service land station set, utilizing the range of wave lengths now generally employed in such receivers.

I have not attempted to describe a receiver which will cure all of the troubles inherent to radio communication, but have simply described one which I believe will certainly add nothing to the complications already existing and which may, and in fact should, without doubt, make it much easier for the operator to secure such results as he secures at the present time with the additional possibilities outlined herein and others which will undoubtedly suggest themselves to other workers.

If thru this paper other companies may take up the subject and determine the full possibilities thru practical application, I shall consider that my work has not been entirely in vain.

SUMMARY: The design and construction of a receiver, operating efficiently and selectively over a long range of wave lengths on any antenna of ordinary dimensions, and controlled by a single handle, are described in detail. The addition of a motor for driving the wave-changing adjustment continuously is shown.

The possibilities of such a receiver for the solution of the interference *problem* are discussed.

DISCUSSION

John V. L. Hogan: Mr. Thompson's admirable paper has brought out clearly the fallacy which underlies our present law on radio signaling. The combination of first requiring a certain definiteness of tuning to minimize interference, and thereafter forcing all public service communication to proceed upon a single wave length is, from the commercial viewpoint, worthy of Alice in Wonderland. From the opposite side, however, it should be noted that by confining the business of radio to a narrow zone of wave lengths and restricting the breadth of tuning so that little interference can be produced outside of that zone, the activities of the commercial companies are effectively limited while the governmental stations are virtually unhampered.

The limitation of radio apparatus, to which Mr. Thompson refers as a basis for the present law, surely need not exist. He proposes that all stations be permitted a substantially free choice of transmitted wave lengths, and that each be equipped with a receiver which will "sweep the ether" with sufficient rapidity to intercept any and all calls, upon whatever wave they may be uttered. This resolves itself into two main elements: (1) a receiver which will respond to successively different wave lengths and which is controlled by a single motion, and (2) a mechanism to adjust the receiver continuously and repeatedly thru its wave-frequency range at a definite rate.

I cannot endorse too heartily the value of the uni-control feature in either a receiver or a transmitter. In my experience with the conditions of commercial and military radio operations I have found a well-tuned receiver to be the exception, whenever primary and secondary circuits were separately adjustable. Frequently the difficulty of establishing or maintaining communication has been traced to nothing more than a failure to tune in a proper manner both circuits of the receiving apparatus. As the International Radio Telegraph Company and its predecessor, the National Electric Signaling Company, have long realized the use of a single variable element for adjusting, the entire receiver will eliminate much of this trouble. I may say, without going into further detail, that we have built simple apparatus embodying this single-variable or uni-control feature, and that our demonstrations have convinced us of its great utility. With it we have invariably been able to pick up signals of unknown wave length in a time much shorter than was possible with the separately adjustable circuits, and frequently we

have discovered and copied messages arriving thru interference which obliterated them when broadly tuned or closely coupled secondary circuits were used. I feel that the single variable tuner should come into extensive use, and that thru it we will be able to solve many of our traffic problems.

As to the second factor, I do not feel so sanguine. It requires a definite time, of say five seconds or more, to recognize a station call in Morse, and at least this interval should be spent in listening on each wave length within the range of the receiver. If we should use the 48 wave lengths proposed by Mr. Thompson, it would require four minutes to complete the cycle and to begin again; manifestly we should be likely to miss many important signals on the momentarily silent or detuned wave lengths. By reducing the number of waves used this difficulty would be correspondingly diminished in importance, but at the same time the possibilities of sinultaneous signaling would be cut down in proportion. I still incline toward the plan of using a specified wave length for all calls and distress signals*, and, by the utilization of single-control senders and receivers, passing to other free but officially determined wave lengths for the transmission of messages immediately after communication has been established. Since all receivers and transmitters remain normally at the "calling" frequency, and since traffic is entirely on other frequencies, this plan should give the maximum protection to life and property at sea as well as the maximum freedom for simultaneous transmission of messages by radio.

Mr. Thompson's receiver appears to vary the tuning of only the antenna circuit. It seems that this process will not give the greatest signal intensity combined with the greatest selectivity. I am sure that all of use will await with interest Mr. Thompson's future paper, which he hints will reconcile the conflicts that, in our present views, limit selectivity of spark signals by the rate at which energy is drawn from the tuned receiver circuits. On the whole, however, we should be gratified by the progress which is indicated in the design of the apparatus described in the paper, and we should join in congratulating Mr. Thompson upon his work.

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*Compare "International Radio-Telegraph Congress," Hogan, "Electrical World," New York, June 22, 1912.

ON THE THEORY OF RADIOTELEGRAPHIC AND RADIOTELEPHONIC RECEIVER CIRCUITS*

By
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In two previous papers (see: "Jahrbuch der drahtlosen Telegraphie und Telephonie," 1909, volume 2, number 6, page 603 and volume 3, number 3, page 302) I have already published the general conditions which will give the greatest efficiency for a receiving set, when the detector (a bolometer, for instance) is inserted in the secondary oscillating circuit. The first object of the present publication is to extend these conditions to receiving sets in which, as is more usually the case, the detector D (a valve of any kind) is (Figure 1) connected across the terminals of the secondary tuning condenser C_2, and a condenser K of relatively high capacity is shunted across the telephone T. A further ob-

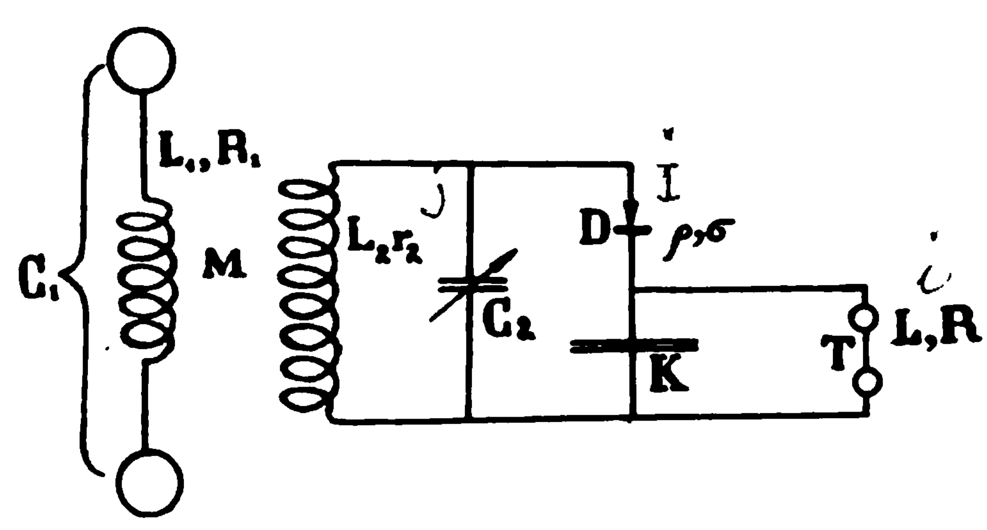

Figure 1

ject is to present some formulas concerning the best adjustment of capacity K and of the constants of the telephone T, especially in case of a beat reception. In this connection, the theory of the "approximate rectifier" (see B. Liebowitz, "Quantitative Relations in Detector Circuits," Proceedings of The Institute of Radio Engineers, volume 5, number 1, page 33, 1917) is

* Received by the Editor, December 3, 1918.

established in a novel manner. For greater clearness, the latter subject will be treated partially in the first portion of this paper, this serving as an introduction to the remaining portions.

I. The Approximate Rectifier

Instead of presenting the characteristic of an approximate rectifier by:

$$I = a_1 V + a_2 V^2 + a_3 V^3 + \cdot \cdot \cdot,$$

as proposed by H. Brandes ("Elektrotech. Zeitschr.," 1906, page 1,015), Tissot, and Liebowitz (former citation), we prefer to employ the power series:

$$V = \rho I + \sigma I^2 + \tau I^3 + \cdot \cdot \cdot, \tag{1}$$

where the current and the voltage are denoted by I and V; and the magnitude and the sign of the coefficients ρ, σ, τ are determined by the shape of the characteristic.

As a rough approximation, we limit the power series (1) to the term of second degree:

$$V = \rho I - \sigma I^2, \tag{1'}$$

and this term will be taken as negative, as shown by a comparison with the first method of expansion, in which the coefficient a_2 is positive (Liebowitz, former citation).

1—Suppose now that at time t, the current thru the condenser C_2 is J, and the current thru the telephone T is i.

Keeping in mind equation (1′), the Ohm and Kirchhoff laws give the following equation:

$$0 = r_2 \int (I + J)\, dt + \rho \int I\, dt - \sigma \int I^2\, dt + R \int i\, dt,$$

since the average value of the emf. induced by the primary winding is zero, if this emf. is periodic (sustained or weakly damped waves) and if the limits of integration are sufficiently separated. But, under the same conditions, we may write:

$$\int J\, dt = 0,$$

and $\qquad$ (2)

$$\int i\, dt = \int I\, dt$$

which means that the average currents thru the condensers are both zero.

It follows that:

$$\int i\,dt = \int I\,dt = \frac{\sigma}{r_2+\rho+R}\int I^2\,dt; \tag{3}$$

consequently, the *average current* thru the detector D (or the telephone) is always different from zero, as well as the square of the effective current.

This result illustrates clearly the rectifying property of the detector. Further, if we employ an electromagnetic (polarised) ammeter and a hot wire apparatus, the sensibility of which is sufficiently high, formula (3) will easily permit the experimental determination of the coefficients ρ and σ. Denote by y the ratio $\frac{\int I^2\,dt}{\int I\,dt}$. The variation of the sum $R+r$, being obtained by means of additional resistance, plot a curve (Figure 2) with different values of y as ordinates and the correspondent values of this sum as abscissas. The curve thus determined from (3) is a

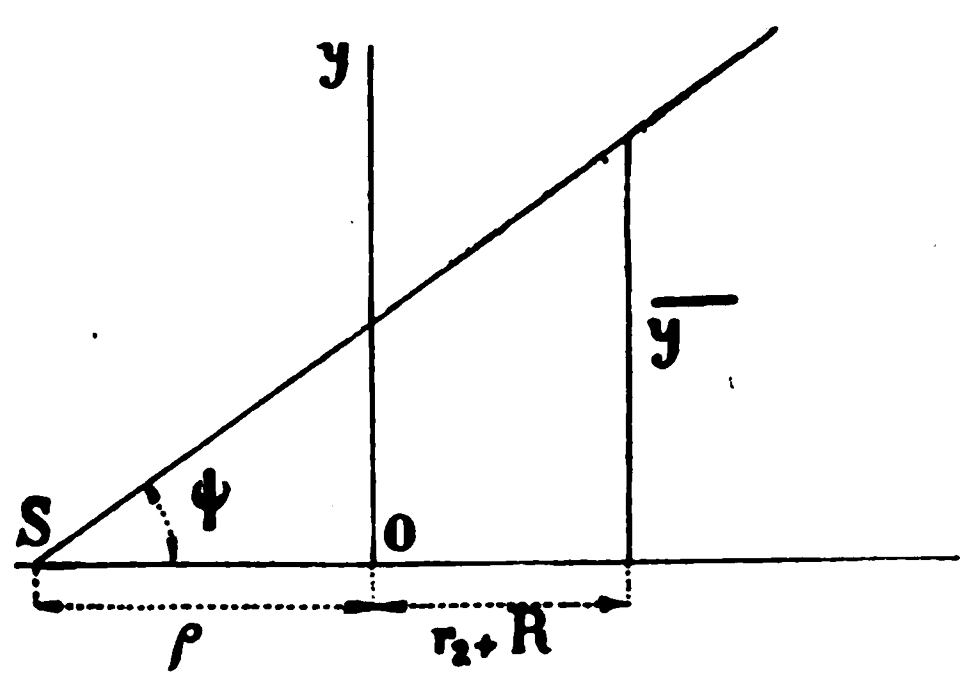

FIGURE 2

straight line, and it cuts the horizontal axis at the point S, the distance of which from the origin O is just equal to ρ; the coefficient σ is given then by

$$\sigma = \frac{1}{\tan\psi}$$

Since the relation (1′) is only approximate, the experimentally-determined curve may differ from a straight line, and it is thus possible to obtain in each case the degree of approximation.

2—Put

$$I = I_2 \sin\Omega t + I_2' \sin\Omega' t + i';$$

the equation (1′) then becomes:

$$V=\rho I_2 \sin \Omega t+\rho I_2' \sin \Omega' t+\rho i'$$
$$-\sigma i' [2 I_2 \sin \Omega t+2 I_2' \sin \Omega' t+i'] \qquad (4)$$
$$-\sigma (I_2 \sin \Omega t+I_2' \sin \Omega' t)^2.$$

In this expression, we may neglect the term which contains $\sigma i'$ as a factor, if both these quantities are sufficiently small.[1] The application of Ohm and Kirchhoff laws to the secondary circuits of Figure 1 thus proves easily that the current

$$I_2 \sin \Omega t+I_2' \sin \Omega' t$$

may be considered as produced by an emf.

$$E_2 \sin (\Omega t+\chi)+E_2' \sin (\Omega' t+\chi'),$$

induced in the secondary winding, and that the current i' flows simultaneously as the result of an emf.[2]

$$e=\sigma (I_2 \sin \Omega t+I_2' \sin \Omega' t)^2, \qquad (5)$$

which originates in the detector D, regarded as a generator of internal ohmic resistance ρ. Of course, according to relation (4), this detector acts as an ohmic resistance of this same value ρ so far as the flow of the current $I_2 \sin \Omega t+I_2' \sin \Omega' t$ is concerned. Finally, the emf. $E_2 \sin (\Omega t+\chi)$ will be taken as the emf. of radio frequency $\frac{\Omega}{2\pi}$ induced from the primary winding, while the emf. $E_2' \sin (\Omega' t+\chi')$ is furnished by a local generator (not indicated in Figure 1), the frequency of which $\frac{\Omega'}{2\pi}$ is given by

$$\Omega'=(1+\varepsilon)\,\Omega,$$

where ε is a small fraction.

This local generator is the "heterodyne" oscillator of the receiving set, and when $I_2'=0$, we obtain the case of the radio-telephone receiver. We can now determine the best conditions for the use of a given detector.

II. The Radio Frequency Circuits

According to the above mentioned results, we may write, using the customary complex method and taking $j=\sqrt{-1}$:

$$E l=R_1 I_1+\left(L_1 \Omega-\frac{1}{C_1\Omega}\right) I_1 j+M \Omega (I_2+J) j,$$
$$0=r_2 (I_2+J)+L_2 \Omega (I_2+J) j+M \Omega I_1 j+\rho I_2, \qquad (6)$$
$$-\frac{J}{C_2\Omega} j=\rho I_2.$$

[1] This assumption will be further justified.
[2] See M. Latour, "Electrical World," April 24, 1915.

In these equations we denote by:

E, the emf. induced by the waves, per centimeter of length of the antenna;

l, the effective length of the antenna;

R_1, L_1, and C_1, the resistance, inductance, and capacity of the primary circuit (including the constants of the antenna);

I_1, the primary current;

M, the coefficient of mutual induction between the two circuits; the other symbols having the same significance as above.

The capacity K is supposed sufficiently large to prevent the flow of any radio frequency current thru the telephone T. It follows that:

$$\begin{aligned} E\,l &= R_1 I_1 + \left(L_1 \Omega - \frac{1}{C_1 \Omega}\right) I_1 j + M\,\Omega\, I_2 j - M C_2 \Omega^2 \rho\, I_2, \\ 0 &= M\,\Omega\, I_1 j + [r_2 + \rho\,(1 - L_2 C_2 \Omega^2)]\, I_2 + \Omega\,(L_2 + r_2 C_2 \rho)\, I_2 j. \end{aligned} \qquad (7)$$

In order to find the conditions for which the maximum current I_2 is obtained, it is possible to employ the method published in the "Jahrbuch." But we prefer to utilize a general theorem which is a very useful one for solving such a problem. Let ϕ_1 be the phase displacement between E and I_1; from the principle of conservation of energy, and taking account of the third of the equations (6), we have:

$$E\,l\,I_1 \cos\phi_1 = R_1 I_1^2 + r_2\,(I_2^2 + J^2) + \rho\, I_2^2 \text{ (effective values),}$$

or

$$E\,l\,I_1 \cos\phi_1 = R_1 I_1^2 + [r_2\,(1 + C_2^2\,\Omega^2 \rho^2) + \rho]\, I_2^2.$$

But we may always write:

$$I_2 = X\, I_1,$$

where the coefficient X is a function of the constants C_1, L_1, M, L_2 · · ·, and so on. We obtain finally

$$I_2 = \frac{X\,E\,l \cos\phi_1}{R_1 + [r_2\,(1 + C_2^2\,\Omega^2 \rho^2) + \rho]\, X^2}. \qquad (8)$$

The angle ϕ_1 is a function of the various constants C_1, L_1, M, L_2 . . . , and as the number of these constants is sufficient, we can choose ϕ_1 and X as independent variables, for a given value of C_2. The maximum of I_2 occurs then when:

$$\cos\phi_1 = 1 \qquad (9)$$

and

$$X^2 = \frac{R_1}{r_2\,(1 + C_2^2\,\Omega^2 \rho^2) + \rho}. \qquad (10)$$

As r_2 is generally very small, we may write approximately:

$$X^2 \doteq \frac{R_1}{\rho}. \tag{10'}$$

From (8) and (9) we get

$$I_1 = \frac{E\,l}{2\,R_1} \cdot \quad \text{(effective values)} \tag{11}$$

Hence, when the current I_2 is a maximum, the primary current is in phase with the impressed emf. and the primary losses are equal to the power delivered to the secondary circuits. When applied to the first of the equations (7), this gives immediately:

$$R_1 I_1 = \left(L_1 \Omega - \frac{1}{C_1 \Omega}\right) I_1 j + M\,\Omega\,I_2 j - M\,C_2\,\Omega^2\,\rho\,I_2, \tag{7'}$$

and therefore, eliminating the ratio $\frac{I_1}{I_2}$, we obtain the two following conditions:

$$R_1 \left[r_2 + \rho\,(1 - L_2 C_2 \Omega^2)\right] + \Omega\,(L_2 + r_2 C_2 \rho)\left(L_1 \Omega - \frac{1}{C_1 \Omega}\right) = M^2\,\Omega^2,$$

$$R_1\,\Omega\,(L_2 + r_2 C_2 \rho) - \left(L_1 \Omega - \frac{1}{C_1 \Omega}\right)\left[r_2 + \rho\,(1 - L_2 C_2 \Omega^2)\right] = M^2 \Omega^3 C_2\,\rho. \tag{12}$$

Discussion: We shall now briefly discuss these conditions:

(a)—From (8), (9), (10), and (11) the maximum secondary current is:

$$I_{2\,max} = \frac{E\,l}{2\sqrt{R_1} \cdot \sqrt{r_2\,(1 + C_2^2\,\Omega^2\,\rho^2) + \rho}} \tag{8'}$$

and the radio power supplied to the detector

$$W_{2\,max} = \frac{E^2\,l^2\,\rho}{4\,R_1\left[r_2\,(1 + C_2^2\,\Omega^2\,\rho^2) + \rho\right]}.$$

This also increases when C_2 decreases, and the limit is

$$\frac{E^2\,l^2\,\rho}{4\,R_1\,(r_2 + \rho)}.$$

(b)—When $C_2 = 0$ (aperiodic receiver) the formulas (12) are reduced to:

$$R_1\,(r_2 + \rho) + \Omega L_2 \left(L_1 \Omega - \frac{1}{C_1 \Omega}\right) = M^2\,\Omega^2,$$

$$\frac{L_1 \Omega - \frac{1}{C_1 \Omega}}{R_1} = \frac{L_2\,\Omega}{r_2 + C}, \tag{13}$$

a result which agrees completely with the formulas of my precious papers, derived by quite a different method.

(c)—It is impossible to assume simultaneously:

$$\Omega^2 L_2 C_2 = \Omega^2 L_1 C_1 = 1,$$

because of conditions (12).

(d)—Eliminating M between these conditions, we obtain a relation which will be fulfilled by the constants L_1, C_1, L_2, C_2, before coupling.

III. The Audio Frequency Circuits

The formula (5) can be written

$$e = \frac{\sigma}{2}(I_2^2 + I_2'^2) - \frac{\sigma}{2} I_2^2 \cos 2\,\Omega t - \frac{\sigma}{2} I_2'^2 \cos 2(1+\varepsilon)\Omega\, t \\ -\sigma I_2 I_2' \cos(2+\varepsilon)\,\Omega t + \sigma I_2' I_2 \cos \varepsilon\,\Omega t. \tag{5'}$$

We retain only the first and the last terms, because the others cannot produce any sensible current thru the telephone T, as the correspondent frequency is too high.

Thus, the emf. e may be considered as the sum of a direct emf.

$$e_d = \frac{\sigma}{2}(I_2^2 + I_2'^2),$$

and furthermore of a sinusoidal emf. equal to:[3]

$$e_a = \sigma I_2' I_2 \cos \omega t;$$

where the frequency $\frac{\omega}{2\pi} = \frac{\varepsilon\,\Omega}{2\pi}$ will be chosen sufficiently low to give a musical tone in the telephone (beat detector).

We will consider now two cases:

(a)—$I_2' = 0$. In this case, the direct current which flows thru the telephone (or polarised galvanometer of any kind), according to the preceding considerations, is:

$$i_d = \frac{\sigma I_2^2}{2(r_2 + \rho + R)}, \tag{14}$$

and as the ampere-turns are proportional to product $i_d\sqrt{R}$, the effect is a maximum when

$$R = r_2 + \rho; \tag{15}$$

[3] This expression, of course, is valuable as long as the approximation (1′) is admissible; if this is not the case, the computations are very complex. Further, when the coefficient σ is sufficiently small, the products by this factor of the currents corresponding to the emf. e_d and e_a will be negligible, and the approximation which leads to the relation (5) is then justified.

This condition determines the number of turns of the winding, when its volume is given.

(b)—$I_2' > 0$. This is the case of the best reception. We will suppose that the reactance ωL_2 and the susceptance ωC_2 are negligible, and write (where $j = \sqrt{-1}$):

$$e_a = (r_2 + \rho) i_a' + L \omega i_a j + R i_a,$$
$$-\frac{i_a' - i_a}{K \omega} j = L \omega i_a j + R i_a.$$

Eliminating i_a', we get:

$$\frac{e_a}{i_a} = R + (r_2 + \rho)(1 - \omega^2 L K) + [L \omega + (r_2 + \rho) K R \omega] j, \qquad (16)$$

a relation which can easily be discussed by means of the diagram of the Figure 3, in which:

$$\overline{OP} = r_2 + \rho + R,$$
$$\overline{PQ} = L \omega,$$
$$\overline{HO} = (r_2 + \rho) L K \omega^2,$$
$$\overline{HA} = (r_2 + \rho) K R \omega,$$
$$\tan \delta = \frac{L \omega}{R},$$

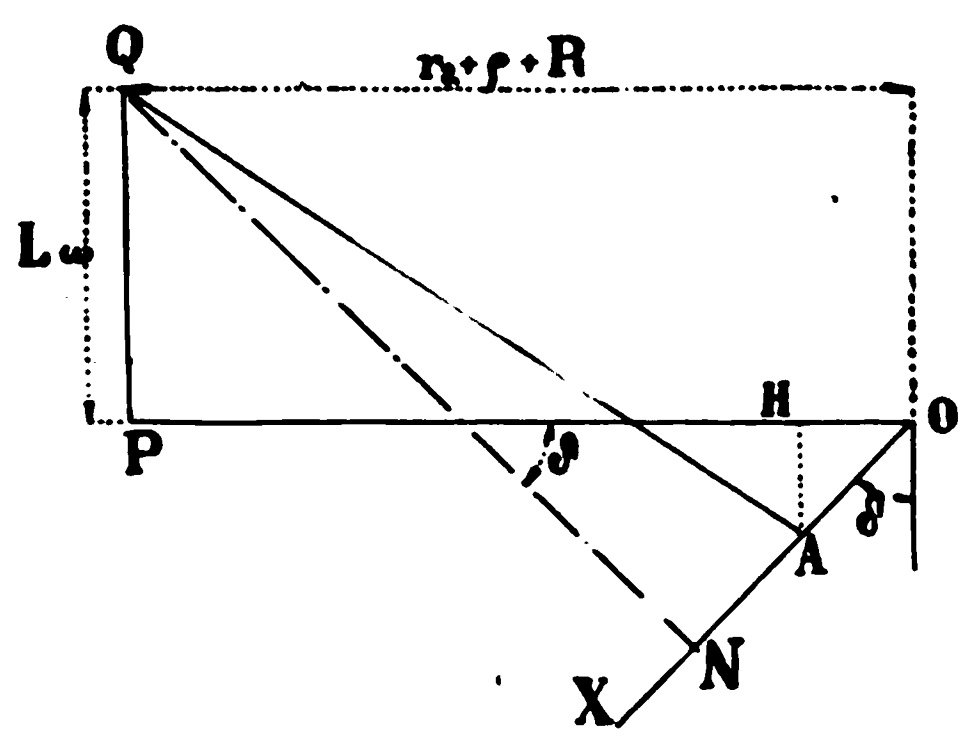

Figure 3

and, therefore,

$$\overline{QA} = \frac{e_a}{i_a}.$$

From (16) we see that the audio frequency current i_a is proportional to the product of the secondary current I_2 by the

current I_2' induced by the local generator (heterodyne), as has been shown for the first time by Mr. Marius Latour (former citation).

If we then change the value of the capacity K, the locus of the point A is the straight line OX, and the maximum of i_a for a given value of e_a, occurs when the point A reaches the point N, QN being perpendicular to OX. It follows from an inspection of Figure 3:

$$\overline{OP} = \frac{\overline{QP}}{\tan\delta} + \frac{\overline{ON}}{\sin\delta},$$

or

$$K = \frac{L}{R^2 + L^2\omega^2} = \frac{\sin\delta\cos\delta}{\omega R} = \frac{\sin^2\delta}{\omega^2 L}. \tag{17}$$

Consequently, the best capacity is independent of the constants of the detector.

The current i_a is then:

$$i_a = \frac{e_a}{\sqrt{[R + (r_2+\rho)\cos^2\delta]^2 + [R\tan\delta + (r_2+\rho)\sin\delta\cos\delta]^2}}. \tag{18}$$

The effective ampere-turns are proportional to the product $i_a\sqrt{R}$, and the ratio $\frac{L\omega}{R} = \tan\delta$ may be supposed to be constant.

The above product is therefore a maximum when:

$$R = (r_2+\rho)\cos^2\delta. \tag{19}$$

It is interesting to compare conditions (15) and (19) which may differ considerably from each other in practice.

It may be further mentioned that the adjustment of the number of turns of the winding can be avoided by means of an audio frequency transformer, inserted between the condenser K and the telephone T, as proposed originally by the late Mr. Jégou.[4] The transformation ratio a of this transformer will be easily adjusted to the best value

$$a \doteq \sqrt{\frac{R}{r_2+\rho}} \times \frac{1}{\cos\delta},$$

according to the previous formula.

The use of this transformer also avoids the flow of the direct current i_d thru the telephones.

SUMMARY: Proceeding from a theory of the approximate rectifying detector, the most advantageous proportioning of the constants of the secondary circuit of a receiver is obtained. The constants of the most desirable telephone winding and the value of the most suitable telephone shunting condenser are then derived.

[4] Mr. Jégou was a radiotelegraphic expert with the Army of the Orient. He died at Florina Hospital in 1917.

DETERMINATION OF RATE OF DE-IONISATION OF ELECTRIC ARC VAPOR*

By

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A direct current arc in a vacuum is maintained by the ionising effect of the current. When the current is stopped for an instant the gas between the arc electrodes becomes de-ionised. The rate of this de-ionisation can be determined by the method here outlined provided certain quantities can be measured with sufficient accuracy.

At the instant the current thru the arc reaches zero the cathode is incandescent and the gas between the electrodes is ionised. The rate at which the gas becomes de-ionised depends upon the ionising effect of the incandescent cathode and upon the rate that the ions disappear from the gas.

When the arc has been extinguished the potential required to re-ignite the arc depends upon the time that the arc has been extinguished. The re-ignition potential also depends upon the kind and pressure of the gas surrounding the electrodes. The kind of gas is largely determined by the material of the electrodes, especially the cathode. According to Dr. Steinmetz the mercury arc in a good vacuum may not restart when extinguished for a few micro-seconds.

Figure 1 shows an arrangement for determining the rate of de-ionisation.

A current I flows thru the circuit $B_1 - L_g - r_o - V_1$ where B_1 is a source of direct current, L_g is the inductance of the circuit, r_o the resistance of the circuit and V_1 is the valve to be tested. The battery B_3 and switch SW are used to prime the valve V_1, and SW is then opened.

Close the switch S-1. This closes the circuit $B_2 - L - R - V_1 - C$ where B_2 is a source of direct current, L is an inductance which is small compared to L_g, R is a resistance, and C is a condenser of capacitance C. Current will flow from B_2 to C until the latter

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is charged to a potential $E_i - E_v$, where E_i is the potential of B_2 and E_v is the drop thru V_1.

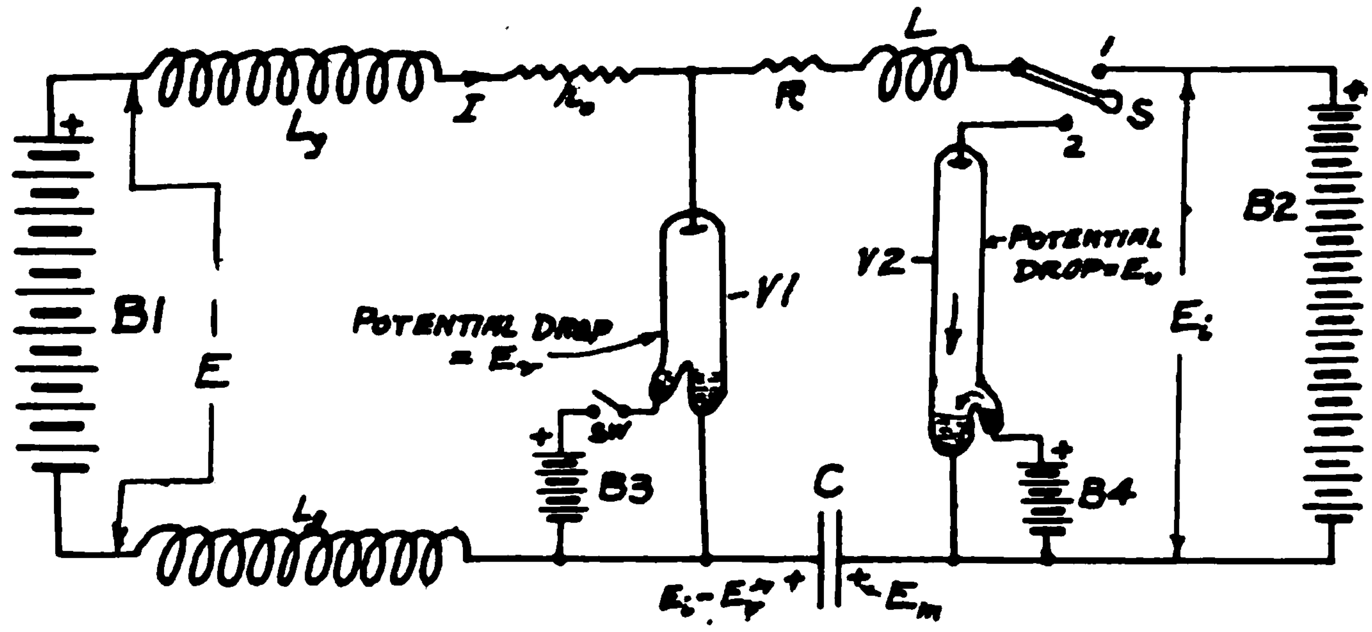

FIGURE 1

Transfer S-1 to S-2 quickly. If the value of $E_i - E_v$ is less than 350 volts, then no spark will pass until S is less than 5x10^{-4} cm. from 2 (See J. J. Thomson's "Conduction of Electricity thru Gases," 1st. edition, page 361). Let S move toward 2 at the rate of 10^3 cm. per second; then the time required to pass from the sparking potential to actual metallic contact between S and 2 will be less than 5x10^{-7} second.

The valve V_2 is a constantly primed mercury vapor valve the sparking potential of which to inverse current is greater than the sparking potential of valve V_1 which is to be tested.

Assuming E_i sufficiently large, the effective discharge of C_1 thru V_1, R, L, and V_2 will momentarily extinguish the arc of valve V_1 and charge C in the opposite sense to a potential E_a, at which the arc is re-ignited. The potential of C increases to E_m, at which instant all the current passes thru the arc again.

Figure 2 shows graphically the potential and current relation of C. The condenser C is part of two circuits; the discharge circuit, $C - V_1 - L - V_2$, and the charging circuit $B_1 - L_g - r_o - L - V_2 - C$.

The extinction and re-ignition of the arc as described above consists of three partial oscillations. The first partial oscillation takes place in the discharge circuit during the interval from the instant of initial discharge of C to extinction of the arc. The second partial oscillation takes place in the charging circuit during the interval of extinction and re-ignition of the arc. The third partial oscillation takes place in the discharge circuit during

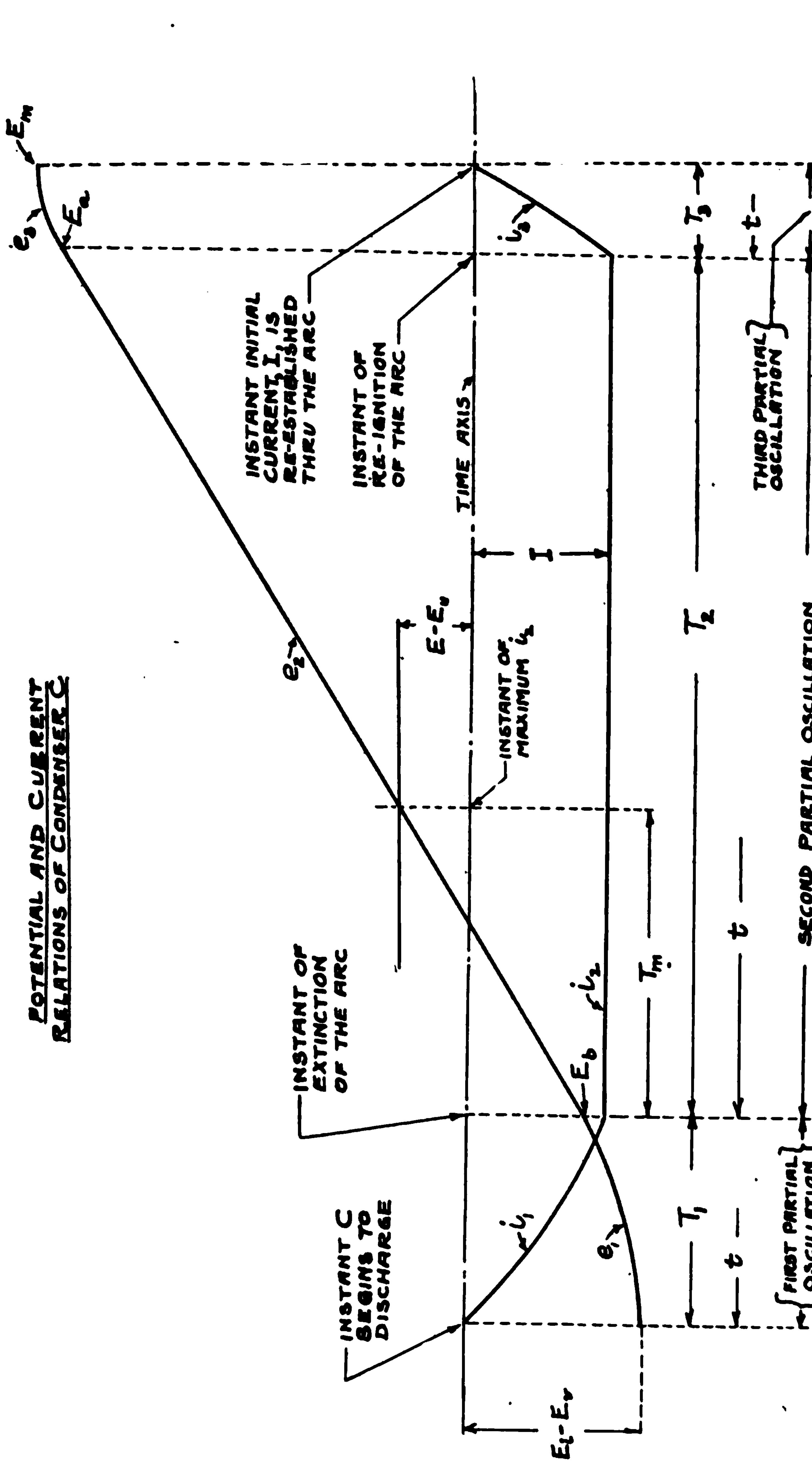
POTENTIAL AND CURRENT
RELATIONS OF CONDENSER C
INSTANT C BEGINS TO DISCHARGE
INSTANT OF EXTINCTION OF THE ARC
INSTANT OF MAXIMUM i_2
INSTANT OF RE-IGNITION OF THE ARC
INSTANT INITIAL CURRENT, I_1 IS RE-ESTABLISHED THRU THE ARC
TIME AXIS
E_m
E_a
E_b
e_1
e_2
e_3
i_1
i_2
i_3
$E-E_0$
E_1-E_0
I
T_m
T_1
T_2
T_3
t
FIRST PARTIAL OSCILLATION
SECOND PARTIAL OSCILLATION
THIRD PARTIAL OSCILLATION

FIGURE 2

the interval from re-ignition of the arc until the current thru V_2 reaches zero.

The ratio between the re-ignition potential E_1 and the duration of the second partial oscillation is a measure of the rate of de-ionisation of the arc vapor. Other quantities remaining constant, a decrease in the capacitance C will increase the slope of e_2; T_2 will decrease; and E_a, the potential of C at re-ignition, will decrease due to less time for de-ionisation of the arc vapor.

The potential drop thru V_1 and V_2 is nearly constant for different values of current thru the vacuum tubes.

First Partial Oscillation

Let i_1 be the instantaneous current in the discharge circuit during the first partial oscillation.

Let L be the inductance, C be the capacitance, and R be the resistance of the discharge circuit. Equating potentials in the discharge circuit.

$$L\frac{di_1}{dt}-E_u+R\,i_1+E_r+\int\frac{i_1\,dt}{C}=0 \tag{1}$$

The solution of (1) is

$$i_1=I_1\,\varepsilon^{-at}\sin\,(\omega t+\theta) \tag{2}$$

where $a=\frac{R}{2L}$, $\omega=\sqrt{\frac{1}{LC}-\frac{R^2}{4L^2}}$ and I_1 and θ

are constants to be evaluated.

When $t=0$, then $i_1=0$, therefore $\theta=0$ or π. From Figure 2, $\frac{d\,i_1}{d\,t}$ is negative when $t=0$, therefore $\theta=\pi$.

When $t=0$ the potential of C is E_i-E_r. Substitute these initial values in (1) and the equation of initial potentials in the discharge circuit is

$$\left[L\frac{d\,i_1}{d\,t}\right]^{t=0}-E_u+[R\,i_1]^{t=0}+E_r+E_i-E_r=0 \tag{3}$$

or,

$$I_1=\frac{E_i-E_u}{\omega L} \tag{4}$$

When $t=T_1$, then $i_1=-I$. Substitute these terminal values and (4) in (2)

$$-I=\frac{E_i-E_u}{\omega L}\varepsilon^{-aT_1}\sin(\omega T_1+\pi) \tag{5}$$

From (5)

$$T_1=\frac{1}{\omega} sin^{-1} \frac{\omega L I}{(E_i-E_u)\,\varepsilon^{-aT_1}} \qquad (6)$$

from which T_1 can be evaluated by one or two trials.

When $t=I_1$, let the potential of C be E_b. Substitute these terminal values in (1),

$$\left[L\frac{d\,i_1}{d\,t}\right]^{t=T_1} -E_u+[R\,i_1]^{t=T_1}+E_v+E_b=0 \qquad (7)$$

From (4), (5) and (7)

$$E_b=(E_i-E_u)\varepsilon^{-aT_1} cos\,\omega\,T_1+aL\,I-E_v+E_u \qquad (8)$$

From (6) and (8)

$$E_b=\sqrt{(E_i-E_u)^2\,\varepsilon^{-2\,a\,t_1}-(\omega L\,I)^2}+aL\,I-E_v+E_u \qquad (9)$$

The value of a may be determined by discharging condenser C. Let I_1 be less than I. Then the arc will not be extinguished by the discharge of C. Let the initial potential of C be E_i' and let E_r be the potential of C at the end of a half cycle when i_1 becomes zero. Then $\omega\,t=\pi$, which substituted in (1) gives

$$\left[L\frac{d\,i_1}{dt}\right]^{t=\frac{\pi}{\omega}} -E_u+E_v-E_T=0 \qquad (10)$$

Equation (10) reduces to

$$(E_i'-E_u)\varepsilon^{-\frac{\delta}{2}}-E_u+E_v-E_T=0 \qquad (11)$$

where $\delta=\dfrac{2\,\pi\,a}{\omega}$.

From (11)

$$\delta=\frac{2}{log\,\varepsilon}\cdot log\frac{E_i'-E_u}{E_T+E_u-E_v} \qquad (12)$$

Second Partial Oscillation

At the instant when the arc is extinguished the direct current circuit becomes the charging circuit which is an oscillating current circuit. The condenser C is charged initially to a negative potential E_b, which is the first partial oscillation terminal potential of C, and its value is expressed by equation (9). Another initial condition of the second partial oscillation is a negative current I flowing in the circuit. Let E_a be the potential of C when V_1 is re-ignited. If the line inductance L_0 is large compared to L then the re-ignition potential of V_1 will be practically E_a+E_u+IR. The capacitance C must be small enough to make the

rise of potential of C rapid and approach a linear function of the time as shown in Figure 2. The current is theoretically a maximum when $t=T_m$.

In the charging circuit let r_2 = resistance, L_2 = inductance, and C = capacitance.

Let $a_2=\frac{r_2}{2L_2}$ and $\omega_2=\sqrt{\frac{1}{L_2C}-\frac{r_2^2}{4L_2^2}}$. The inductance $L_2=L_g+L$ and resistance $r_2=r_o+R$.

The equation of potentials in the discharge circuit is

$$E+L_2\frac{d\,i_2}{d\,t}+r_2\,i_2-E_u+\int\frac{i_2\,d\,t}{C}=0 \tag{13}$$

The solution of (13) is

$$i_2=I_2\,\varepsilon^{-a_2t}\,sin\,(\omega_2\,t-\beta) \tag{14}$$

where I_2 and β are constants to be evaluated.

Referring to T_m of Figure 2, $\beta=\omega\,T_m+\frac{\pi}{2}$.

When $t=0$ then $i_2=-I$. Substitute these initial values in (14).

$$I=I_2\,sin\,\beta \tag{15}$$

When $t=0$, the potential of C is E_b. Substitute these values in (13) to get the equation of initial potentials which is

$$E+\left[L_2\frac{d\,i_2}{d\,t}\right]^{t=0}+[r_2\,i_2]^{t=0}-E_u+E_b=0 \tag{16}$$

From (15) and (16)

$$E-\omega_2L_2\sqrt{I_2^2-I^2}+a_2L_2\,I-r_2\,I-E_u+E_b=0 \tag{17}$$

From (17)

$$I_2=\frac{1}{\omega_2L_2}\sqrt{(E_b+E-E_u-a_2L_2I)^2+(\omega_2L_2I)^2} \tag{18}$$

From (15) and (18)

$$\beta=tan^{-1}\frac{\omega_2L_2\,I}{E_b+E-E_u-a_2L_2\,I} \tag{19}$$

When the current i_2 charges C to a potential E_a, the arc V_1 is re-ignited. The value of E_a depends upon the time T_2 between extinction and re-ignition of V_1 and upon C, I_R and E_b. At re-ignition the following terminal potential relations exist in the charging circuit.

$$E+\left[L_2\frac{d\,i_2}{d\,t}\right]^{t=T_2}+[r_2\,i_2]^{t=T_2}-E_u-E_a=0 \tag{20}$$

which reduces to

$$E+\varepsilon^{-a_2 T_2}\left\{\left[\frac{I}{C}-a_2\left(E_b+E-E_u\right)\right]\frac{\sin\omega_2 T_2}{\omega_2}-\left(E_b+E-E_u\right)\cos\omega_2 T_2\right\}$$
$$-E_a-E_u=0 \qquad (21)$$

Equation (21) expresses a relation between T_2 and E_a.

The value of E_a will be obtained from E_m in the third partial oscillation.

Solving (21) for T_2

$$T_2=\frac{1}{\omega_2}\left[\sin^{-1}\frac{\left(E_a+E_u-E\right)\varepsilon^{a_2 T_2}}{\sqrt{\left[\frac{I}{\omega_2 C}-\frac{a_2}{\omega_2}\left(E_b+E-E_u\right)\right]^2+\left(E_b+E-E_u\right)^2}}\right.$$
$$\left.+\tan^{-1}\frac{E_b+E-E_u}{\frac{I}{\omega_2 C}-\frac{a_2}{\omega_2}\left(E_b+E-E_u\right)}\right] \qquad (22)$$

When a is negligible, (22) reduces to

$$T_2=\frac{1}{\omega_2}\left[\sin^{-1}\frac{\omega_2 C\left(E_a+E_u-E\right)}{\sqrt{I^2+\omega_2^2 C^2\left(E_b+E-E_u\right)^2}}\right.$$
$$\left.+\tan^{-1}\frac{\omega_2 C\left(E_b+E-E_u\right)}{I}\right] \qquad (23)$$

In (21) let $a=0$, let $\sin\omega_2 T_2=\omega_2 T_2$ and let $\cos\omega_2 T_2=1$, then

$$T_2=\frac{C}{I}\left(E_a+E_b\right) \qquad (24)$$

Equation (24) is based upon the assumption that L_g is infinitely large and that I is constant during the interval T_2.

Let the charging current at the instant of re-ignition be $-I'$; then from (14) and (15)

$$-I'=\frac{I\,\varepsilon^{-a_2 T_2}}{\sin\beta}\sin(\omega_2 T_2-\beta) \qquad (25)$$

THIRD PARTIAL OSCILLATION

The third partial oscillation is similar to the first except that the initial potential and current in the circuit is different.

Equations (1) and (2) apply except I_1, and θ will have different values. Call these values I_3 and ϕ then (2) becomes

$$i_3=I_3\,\varepsilon^{-at}\sin\left(\omega t-\phi\right) \qquad (26)$$

When $t=0$ then $i_3=-I'$. Substitute these initial values in (26)

$$I'=I_3\sin\phi \qquad (27)$$

Let E_m be the potential of C when $I_3=0$ and $t=\frac{\phi}{\omega}=T_3$.

Then from (1) the equation of terminal potentials is,

$$\left[L\frac{d\,i_3}{d\,t}\right]^{t=T_3} - E_u + [r\,i_3]^{t=T_3} + E_v - E_m = 0 \tag{28}$$

from which

$$\omega L\,I_3\,\varepsilon^{-aT_3} - E_u + E_v - E_m = 0 \tag{29}$$

From (27) and (29)

$$\phi = \omega\,T_3 = sin^{-1}\frac{\omega L\,I'\,\varepsilon^{-aT_3}}{E_m+E_u-E_v} \tag{30}$$

from which ϕ or T_3 may be evaluated by one or two trials when $a\,T_3$ is not negligible.

Since E_a is the potential of C at re-ignition equation (1) becomes

$$\left[L\frac{d\,i_3}{d\,t}\right]^{t=0} - E_u + [R\,i_3]^{t=0} + E_v - E_a = 0 \tag{31}$$

from which

$$E_a = \omega L\,I_3\cos\phi - aL\,I' - E_u + E_v. \tag{32}$$

From (30) and (32)

$$E_a = \sqrt{(E_m+E_u-E_v)^2\,\varepsilon^{2aT_3} - (\omega L\,I')^2} + E_v - E_u - aL\,I' \tag{33}$$

To evaluate E_a from (33) assume $I'=I$ and solve for E_a. Use this first trial value of E_a in (24) and solve for T_2. Use this first trial value of T_2 in (25) and solve for I'. Use this value of I' in (33) and (30) for a final value of E_a. A practically exact value of T_2 may be obtained by using these values of T_2 and E_a in (22) or (23).

De-ionisation Curve

To show the rate of de-ionisation graphically a curve can be plotted for different values of E_1 and T_2. In order that the conditions for de-ionisation remain constant it is necessary to maintain I, E_u, R, and E_b constant. From (24) is seen that T_2 must be changed by varying C. To keep E_b constant, (9) shows that E_i, ωL, and I must remain constant.

This curve will show the de-ionisation characteristics of an arc under given conditions. To compare an arc under different conditions of electrode material, temperature and pressure the ratio $\frac{E_1}{T_2}$ will indicate the relative rates of de-ionisation when either E_1 or T_2 is assigned a definite value.

Figure 3 shows the general form of a T_2–E_1 curve. As the capacitance C is increased the potential of C rises more slowly while the re-ignition potential E_1 rises. By keeping $E_b - E_u - I\,R$ constant, the inverse potential impressed upon the valve will not affect the value of E_1.

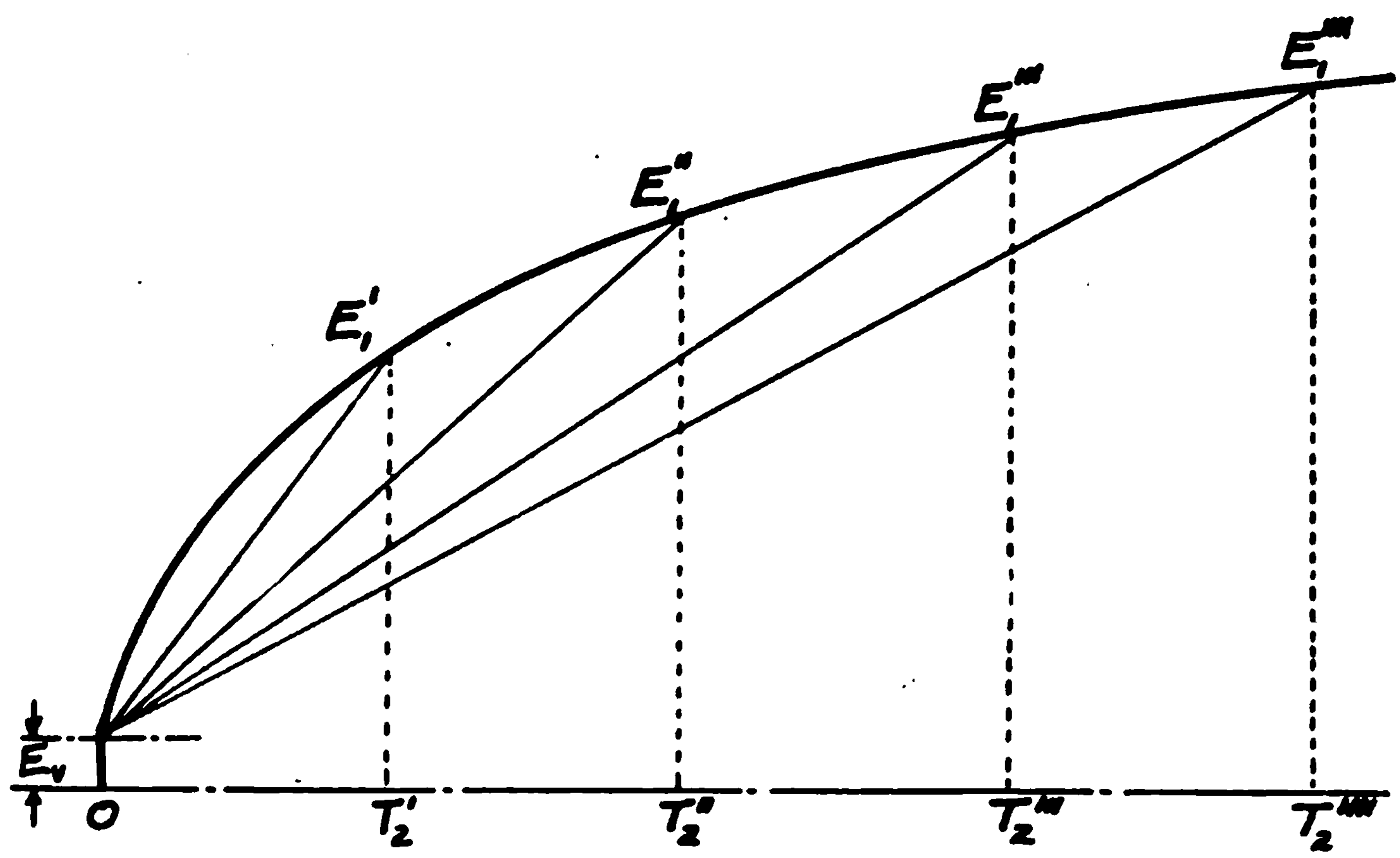

FIGURE 3

Another important curve will be obtained by plotting E_a and E_b in (24) while T_2, I and C remain constant. E_b can be varied by changing the initial potential E_i. The effect upon E_a of changing the inverse potential upon the valve can thus be ascertained.

NUMERICAL EXAMPLE

The application of the preceding equations can be illustrated by assuming values for the quantities which can be measured.

Assume $E_i' = 90$ volts, $E_u = E_v = 20$ volts, and $E_T = 63.4$ volts. From (12), $\delta = 0.2$.

Let $T_n = \dfrac{1}{f}$ = the natural period of the discharge circuit. Assume $f = 10^5$ and $C = 0.2\ \mu$f.

Then $\mathrm{L} = \dfrac{10^6}{8\pi^2}$, $\omega = 2\pi \,.\, 10^5$, $\omega L = \dfrac{10^{11}}{4\pi}$, $a = 0.2 \,.\, 10^5$, $aL = \dfrac{10^{10}}{4\pi^2}$ and $T_n = 10$ micro-sec. Assume $E_i = 110$ volts and $I = 10$ amperes.

From (6), $T_1 = 0.663$ micro-seconds, and from (9) $E_b = 24$ volts. In (30) and (33) assume $I' = I$ and $E_m = 500$ volts. Then

$T_3 = 0.25$ micro-sec., and $E_a = 491.5$ volts. Substituting in (24), $T_2 = 10.3$ micro-seconds. Since $R = 2fL\delta$, $R = 0.5$ ohm and $R = 516.5$ volts.

These values of E_1 and T_2 are based upon the assumption that L_g is infinitely large.

Assume $L_2 = 100L$ and $E = 110$ volts.

From $r_o = \frac{E - E_r}{I}$, $r_o = 9$ ohms. Since $r_2 = r_o + R$, $r_2 = 9.5$ ohms.

The natural period of the charging circuit is then $T_c = 100$ micro-seconds and $\omega_2 = 2\pi \cdot 10^4$.

Substitute this value of ω_2 in (25), then $I' = 7$ amperes. From (30) and (33) $E_a = 495.2$ volts, and $E_1 = 520.2$ volts.

Solving (22) completely, $\omega_2 T_2 = 38.6°$ and $T_2 = 10.7$ micro-seconds. An error of only 4 per cent. was made by assuming L_g to be infinitely large.

The value of L_g must be determined for a natural period of 100 micro-seconds. An iron magnetic circuit will increase the inductance and energy stored in a given coil. During the time T_2, part of this energy is transferred to condenser C and part is dissipated by eddy currents in the iron. Other conditions remaining the same, the ratio of energy transferred to C to energy dissipated by eddy currents in L_g decreases with a decrease in capacitance C. In other words, L_g is variable when its value depends upon the presence of iron. Iron should therefore be used only when L_g is to be considered infinitely large. When no iron is used, the source of the direct current should be a generator shunted with a large condenser or a battery.

The charging current circuit resistance r_o must be so constructed that its value will be the same for a current of period of 100 micro-seconds as for a continuous current.

Starting Terminal

The effect of a starting terminal on an electric arc valve will reduce the re-ignition potential.

Figure 4 is similar to Figure 1 except that V_1 is provided with a starting terminal T the presence of which will reduce the re-ignition potential E_1. The effect of a starting terminal is illustrated by the starting band of a mercury vapor lamp. The potential between T and K is the same as the potential of C. It may be made less by dividing C into a number of condensers in series and connecting T to a point between the condensers.

In Figure 4, the valve V_2 is not a constantly primed valve.

The switch S may be moved slowly from 1 to 2 and a sufficient potential impressed upon C_u for a spark to pass from F to K_2, which will prime V_2 and allow C to discharge. When this method of discharging C is used the potential E_i may be made as high as desired.

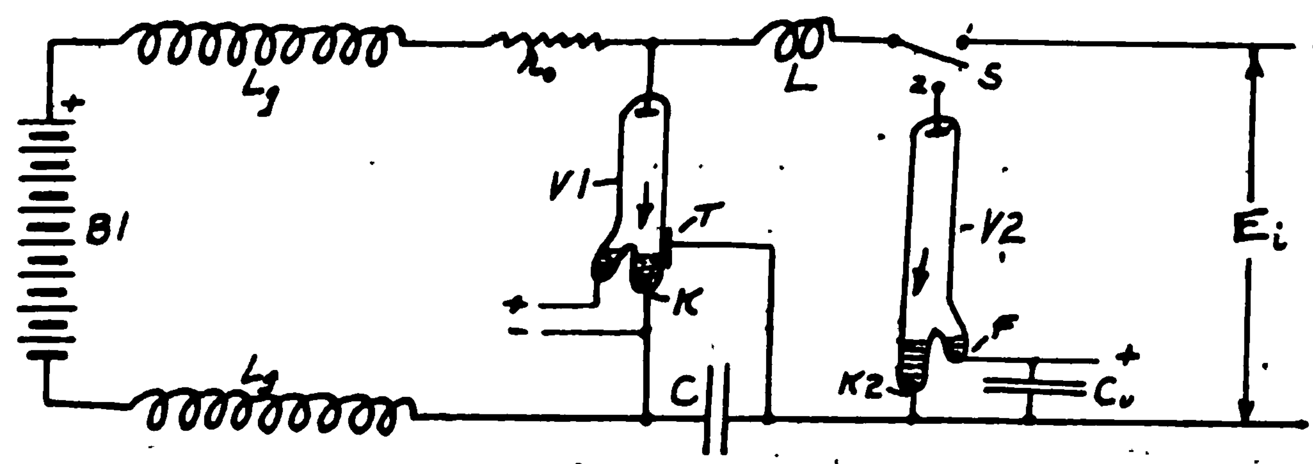

FIGURE 4

The re-ignition potential E_1 may be increased by placing the arc in a magnetic field.. This is illustrated by the Poulsen arc.

Inverse Current

The maximum inverse potential impressed upon V_1 is $E_b - E_u - I\,R$. This potential will produce inverse current which is difficult to measure at high frequencies. The following arrangement is suggested.

Figure 5 is similar to Figure 4 except the additional valves V_3 and V_4 have been introduced and T omitted. All current passing from anode to cathode in V_1 passes thru V_4, but all inverse current thru V_1 passes thru V_3. Place a transient-current-indicating device D in series with V_3. Then D will indicate the inverse current thru V_1. The value of E_u in the equations will be the potential drop thru V_1 and V_4 in series.

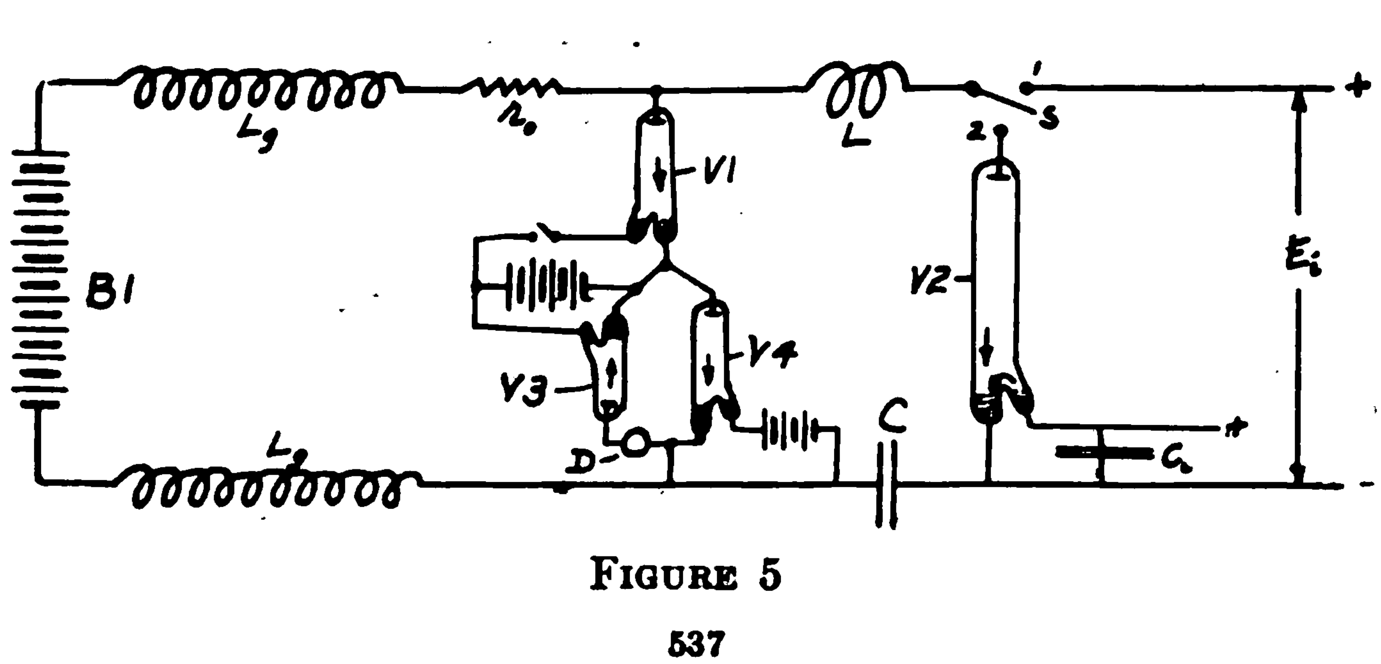

FIGURE 5

It may be possible, with this arrangement to ascertain the cause for the inverse discharge at the beginning of each half cycle in a mercury vapor rectifier. (See "General Electric Review," October, 1913, page 701.)

REMARKS

There is a minimum current which will sustain a given arc. The maximum amplitude of i_1 may be less than I and still the arc will be extinguished. If the arc is not extinguished, the final potential of C will be E_T, which is less than E_i, but when extinction takes place the final potential of C will be E_m, which is greater than E_i. To extinguish the arc the ratio $\sqrt{\frac{L}{C}}$ must be less than the ratio $\frac{E_i-E_u}{I}$, so that the term under the radical in (9) is positive. In order to use a low potential for E_i it is necessary to make L very small when the natural period of the discharge circuit is a micro-second or less. If V_2 in Figure 1 is omitted, V_1 will probably be permanently extinguished.

The relation between E_1 and T_2 for an arc in mercury vapor at low pressure will probably be expressed in hundreds of volts and a fraction of a micro-second because of the known quenching action of mercury vapor.

The equations of the first and second partial oscillations are applicable to the Poulsen arc but in the third partial oscillation the value of E_v varies. Its average value is much higher than in the first partial oscillation. This has been shown with a Braun tube. If E_m is not much greater than E_a, then valuable information may be gained by the method here described.

The accuracy of the results in the numerical example would have been increased if E had been reduced from 110 volts to 30 volts. The effect of r_o would then have been still more negligible.

If E_i-E_u is measured with a ballistic galvanometer, then C may be charged to a potential E_i and then discharged thru the galvanometer and V_2 in series which will give the potential difference directly.

The method here developed is based only upon theoretical deductions. Experimental research is required to determine the rate of de-ionisation of arc vapor under different conditions. The results of such research will be of great value.

SUMMARY: After discussing the de-ionisation and consequent loss of conductivity of mercury vapor carrying a momentary arc, the author considers an arrangement of circuits for determining the rate of de-ionisation, and the effect of this rate on the voltage required for a subsequent re-ignition.

The theory of the circuits shown is given, and illustrated by numerical examples. It is suggested that the arrangements shown be experimentally carried out because of possibly important practical applications.

THIRD DISCUSSION ON

"THE ELECTRICAL OPERATION AND MECHANICAL DESIGN OF AN IMPULSE EXCITATION MULTI-SPARK GROUP RADIO TRANSMITTER

(A Paper by Ensign Bowden Washington, U.S.N.R.F)

By

LIEUTENANT ELLERY W. STONE, U. S. N. R. F.

(OFFICER IN CHARGE, U.S. NAVAL RADIO STATION, SAN DIEGO, CALIFORNIA)

In reply to Mr. Washington's discussion, which reached me this date, I should like to state that I am quite in accord with his definition of impact excitation, as is evidenced by definitions of the term given in my previous articles on the subject. If, however, thru accident or design, the gap circuit of an impulse transmitter delivers a little more than the single half cycle which characterizes its normal operation, it can hardly be placed in the ordinary quenched gap type of transmitter, inasmuch as the excitation of the antenna is still of the shock type. The major portion of the energy is unquestionably resident in the first half cycle.

In connection with the tests of the Kilbourne and Clark transmitter conducted by Mr. Washington at Harvard University in which he obtained two and one half oscillations, it is presumed that these tests were those conducted for the Marconi Company in 1916 during its suit against the Kilbourne and Clark Company. It may be stated that the defendant (Kilbourne and Clark) proved to the satisfaction of the Court that these tests were conducted under abnormal conditions,to-wit; abnormal gap length, and improper adjustment of the circuits, and that attempt was made to produce oscillations in the gap circuit rather than to demonstrate normal operation.

If the photographs showing a single impulse in the gap circuit were not indicative of impact excitation, they were not so considered by the Court. These were taken at the University of Washington by Lieutenant Greaves, who had considerable experience in this work at Harvard University, in the presence of Mr. F. A. Kolster, Mr. Frederick Simpson, myself and several others.

In the decision of the Court, in which the impulse nature of this transmitter was upheld, Judge Netterer rules as follows:

"The defendant, to demonstrate the fact of the 'single chunk'

(impact, E. W. S.) "conversion of antenna energy, introduced the result of experimentation conducted at the University of Washington with the Braun tube on the behavior of the Simpson Mercury Valve Transmitter. . . The deflection of the spot of light across the screen corresponds with the motions of the current, first in one direction and then in the other, and if the disturbing" (gap, E.W.S.) "circuit is not characteristically an oscillating circuit . . . the spot of light would appear as in the photograph. . . The contention of the plaintiff" (Marconi Company) "with relation to the Massachusetts" (Harvard University) "test, in which it was shown that there were two and one half oscillations in the circuit, and that this must refute the contention of the defendants with relation to the Washington University photographic test, may be answered by the suggestion that the Washington University result was obtained,—the photograph speaks for itself—and defendant's witnesses to that extent are corroborated. The Massachusetts experiments show that there were many elements that entered into the experiments with relation to the appliances and the adjustment of the apparatus. Dr. Zenneck's testimony, which does not seem to be denied, shows that photographs were only taken when the adjustments were such as to produce the desired result, and that the effort was for the purpose of obtaining evidence of oscillations in the trigger" (gap, E.W.S.) "circuit, rather than to present to the Court the result of all the experiments that were made, together with the adjustments for each result. . . No facts shown indicate that the oscillations" (antenna) "were the result of resonant transfer of energy."

I do not feel required to defend the term "partial discharge," the expression being taken from Zenneck's writings. I should say that the definition of "full charge" which Mr. Washington seeks, is that charge given the condenser by the time the peak of the transformer secondary wave has been reached, the separation of the gap permitting, or the maximum charge which the condenser can receive with a given secondary potential. This "full charge" will be recognized as distinct from the partial charge which the condenser ordinarily receives as a result of the minute separation of the impulse gaps, thus giving rise to several discharges of the condenser per alternation.

February, 17 1919.

FURTHER DISCUSSION ON "RECEPTION THRU STATIC AND INTERFERENCE"

By

Roy A. Weagant

Lee De Forest (by letter): I regret that I was out of the city when Mr. Weagant's paper on static elimination was presented, and particularly that copy of his paper was not received until the Proceedings were ready to issue.

I have read with intense interest the paper and such discussion as appeared with it. Such abundance of well merited praise has therewith appeared that I feel justified in limiting my present remarks chiefly to helpful criticism. I too, perhaps earlier than most of our readers, have fought with static, and been burned by it. I can honestly state that I know from long and intimate acquaintance what genuine, sub-tropical, summer static is—perhaps far better than some others who from cool northern laboratories have casually announced its final "elimination."

My experience has indeed been such that even now, as during the earlier years, I doubt that the "99 per cent, reliable, guaranteed, copper-riveted static eliminator" has been discovered. And certain recent careful inquiries among some very capable radio observers and experts (some in the Government service) have elicited statements, the accuracy and fairness of which I am bound to accept, to the effect that the Weagant eliminator, like a legion of others, falls down at times. It is, in short, by no means all that its enthusiastic proponents claim.

If such be the fact, I regret it as much as anyone. Having spent all my working years in the development of the radio art, I rejoice at each actual step in advance, by whomsoever wrought. Static has been our *bête-noir* from "Genesis"; and I too await for "Revelation"—I hope to see it. But let us not throw our hats skyward over the assassination of "Satan Static," until further evidence than that of the inventor of an eliminator, and other interested observers, comes forward after long summer months of trial, to lay its wreath of unbiased testimony upon the "Tomb of Trouble."

Possibly I have been misinformed. Let other users of the Weagant system, not employees or associates, who have successfully used the system thru say six summer months of European radio reception in the South, come forward. In The Institute of Radio Engineers we simply want facts, not flattery.

Mr. Weagant's paper can be divided into two parts—The first and by far the largest and most exhaustive, outlines in great detail the amazing lengths and profligate expense to which the Marconi Company has gone to accomplish what Mr. Weagant himself admits in the brief, *second* part of his paper, can be achieved by far simpler and more rational methods. Therefore we may dismiss with sincere praise for the inventor's courage and persistence the consideration of vertical loops 400 by 1,000 feet (120 by 300 m.) in dimensions and located miles apart, or of horizontal antennae 6 miles (9.6 km.) long, of miles of pasteboard tubes covered with tinfoil "gleaming in the Florida moonlight." Sad indeed would appear the future of radio communication if we believed that static elimination depended in the slightest degree on the utilization of devices of such Brobdingnagian dimensions. I do not believe that it is essential to separate the "static tank" loops as one does water tanks along a railroad.

But while describing these experiments Mr. Weagant is sadly misleading in his historical omissions. To those not well versed in early radio history it would have been only instructive and fair had he pointed out that Mr. John Stone Stone first conceived and described the two vertical receiving antennas, separated by a half-wave length, or portion thereof, arranged in the plane of propagation—differentially connected, thru long horizontal leads, to a common receiving system, to balance out interfering disturbances. In Stone's U. S. patent 767,970, filed June, 1901, this system is carefully outlined and analyzed, together with the third vertical antenna located directly at the receiver. (This latter is for use as a transmitter, the arrangement b ing such that the powerful impulses from this transmitter shall be completely neutralized upon the two receiving systems.)

Mr. Weagant mentions "among the early workers with the loop, Bellini—Tosi, and Braun"; but completely ignores the fact, of which we Americans should be proud, that the first disclosure of the loop receiver and direction finder was in Stone's U. S. patent, filed January 23, 1901; and that he laid down at this remote date the basic principles on which the entire art of direction finding (and incidentally of Mr. Weagant's eliminator) are founded. Mr. Weagant further erroneously ascribes the first use of the horizontal linear aerial to Mr. Marconi. As a matter of fact this was, I believe, also an American discovery—at Block Island in 1903.

It seems to be equally unknown that the first patent on the horizontal receiving loop or horizontal conductor, with length independent of the wave length and rotating around a vertical axle, for localizing the direction of incoming signals, was issued to the writer, in 1904.

Mr. Weagant describes the differentially combined "interference-prevention" circuits of Fessenden as "fundamentally incorrect" because "the detuning of one branch circuit affects the intensity of both the signal and static currents in the secondary circuit in the same ratio." As a matter of fact, where enlightened methods of this sort are employed in receiving undamped waves, a very great improvement in the signal-static ratio may be noted.

Moreover, after styling the Fessenden differential principle as "fundamentally incorrect," Mr. Weagant actually relies on exactly that principle to achieve the final elimination of static impulses between his "static tank" circuit and his third static-signal circuit. Note his arrangements in Figures 7, 11, 13, 18, 30, and so on.

The theory Mr. Weagant advances of the vertical origin of the "grinders" is novel, if we consider that he means that all such disturbances come from directly above the receiving station. But the long established fact that the same violent static impulse is frequently noted simultaneously at stations separated by hundreds of miles precludes the acceptance of this view. Then, moreover, would it be possible to eliminate such disturbances by rotating a small loop-receiving antenna about a horizontal axis.

Both audio and radio balancing between two helices at right angles have been tried by Major Charles A. Culver in recent tests for the Signal Corps in an effort to neutralize static effects, but without success. Simultaneous photographic records of "X" impulses received on two helices placed at right angles to one another and in a vertical plane appear to show that the extraneous disturbances do not occur simultaneously at two mutually perpendicular planes.

It has long been accepted generally that static impulses originating in the tropics are reflected by the Heaviside layer, or by upper banks of ionized air, and consequently reach northern and southern latitudes with a downward vertical component. This explanation was cited by Professor Pupin in his discussion of the Weagant paper. Moreover if the startling theory of vertical origin were correct it would obviously become a very

simple matter to neutralize the static impulses on two small vertical loops, placed at right angles, while receiving the desired signals on that loop which lay in the plane of their propagation. Such, of course, is by no means the case. The germ of genuine merit in Mr. Weagant's voluminous paper resides, it seems to me, in the idea of the so-called "static-tank," balancing out the signal in two receiving systems, leaving only static, and then balancing that against the static in a third system, obtaining finally only a residuum of signal, and then amplifying this remnant. But what justifiable considerations lead him to accomplish these effects by use of antenna systems which require an entire township for their exploitation, "across a canal, thru cow pastures and frequently broken?" Certainly it was *not* to avoid the use of the audion amplifier! No such reluctance in employment of another's device to accomplish so useful an end, characteristic tho it be with certain investigators, would be justifiable.

To illustrate how simple it would be to accomplish by sensibly small loop antennas what Mr. Weagant does with his gigantic "loops" connected to his receiver by miles of horizontal conductors (systems which possibly operated not as *true loop* antenna at all) let me cite the results obtained by Latour at Lyons. Using three radio frequency audion amplifiers, in cascade, an audion detector, and then several audio frequency amplifiers, in cascade, the French military administration has been receiving from Annapolis at a point only one mile from the Lyons arc transmitter station—without any interference therefrom whatever,—while Lyons was transmitting at full power (150 kilowatts) with a wave-length only 2.5 per cent different from that of Annapolis. And Latour does this using as a receiving antenna *a coil only* 60 *cm.* (24 *inches*) *in diameter!* Contrast this with one 400 by 1,000 ft. (120 by 300 m.) Of course, French static is not to be compared with our summer variety; but every advantageous result which Mr. Weagant describes can, I believe, be accomplished (as Prof. Pupin urges) with receiving circuits of very ordinary dimensions, perhaps "such as can be used on board ship." For compare these results at Lyons with what the huge Weagant arrangement accomplishes; "The reception of Carnarvon's signal, 14,200 meters, thru the powerful interference of the 200 kilowatt Alexanderson alternator (supposedly perfect sine-wave emission) at New Brunswick, *only* 25 *miles* (40 *km.*) away, working at 13,600 meters (a 4.4 per cent difference), has been an every day performance of the system."

As further bearing out this statement, and Latour's results, consider the results described by Lieut.-Commander A. H. Taylor in the June number of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS—obtained with loops only 3 meters (9 feet) square. If one uses such loops it will not be "found necessary to have an operator located at the remote loop to make adjustments in accordance with instructions telephoned to him by the observer in the receiving station, using the cable wire for this purpose." I must disagree with Mr. Weagant's statement that "very satisfactory practical working was secured," since, to quote further, "with both these arrangements local tuning of the loops was necessary, and this always involved a tedious adjustment until the correct setting for a given wave length was obtained, and even when this setting was known, it was necessary for some one to go to each of the loops—not a convenient procedure with antennas three miles (4.8 km.) apart."

A stiff course in practical "statics" has heretofore been recommended as an effective cure for any tendency to pedantic and dogmatic theorization on their behavior. Hence, after studying Mr. Weagant's Figures 9, 10 and 17, one can but wonder that any person who has struggled with the irrational idiosyncrasies of subtropical static as Mr. Weagant has, can still retain sufficient patience, or religious faith, to attempt to describe their operation in terms of cosine curves, and "azimuthal angles!" In others, therefore, who have likewise gone thru the static mill a moderate amount of more or less profane skepticism as to the pertinence of these figures must be tolerated.

A common oversight on the part of inventors of most so-called "static eliminators" has been the fact that powerful static disturbances may reach the highly sensitive detector-amplifier directly thru the low-frequency conductors attached thereto, that is, the battery leads, the telephone cord, the body of the operator himself, or act directly upon the sedondary coil, or, if unshielded, the metallic plates of the condensers attached to the receiver. Thus no matter how perfect and intricate a "filtering" system may have been installed between the antenna and the detector, the disturbances which have been successfully barred from entrance by the fortified front of the house steal in thru the back, where they find the entire roof and walls missing. To expect then to exclude effectively static disturbances from a receiving and detector system, the various elements of which are not effectively and in succession screened electrically from the other links in the filtering system, is as futile as to hope to per-

form delicate photometric work in a room elaborately equipped with the proper apparatus, but the walls of which are white or silvered, and where on all sides are unmasked arc lights, dazzling sparks, magnesium flares, and windows open to the sunlight. Especially does the above analogy hold good when violent electrical storms are in the neighborhood of a long distance radio receiving station.

From the foregoing, it should be apparent that it is practically futile to conduct radio signaling which shall be at all times immune to static disturbances and interruption so long as it is possible for violent disturbances to produce in the receiving system effects which simulate those produced by the waves emitted from the transmitting station. Yet this situation is inevitable so long as attempts to solve the problem are limited exclusively to the receiving apparatus.

Unquestionably, Mr. Weagant has made a contribution of genuine value towards the long-desired goal. But until efficient shielding methods are adopted there are still certain to occur intervals when the only satisfactory means for cutting out unwelcome disturbances is that which was so effectively adopted during the public discussion of Mr. Weagant's paper—lay the "phones on the table, and refuse to listen."

Roy A. Weagant (by letter): The tone of Dr. de Forest's comment is such as to tempt one to ignore it. Also, to one familiar with the paper, his failure to read it carefully is so evident that a reply seems unnecessary. However, lest silence should seem to give assent to some of his misinterpretations, the following reply is submitted.

Dr. de Forest has expressed doubt that the "99 per cent reliable, guaranteed, copper-riveted, static eliminator" has been discovered. Reference to the text of the paper fails to disclose any statements in this language, or which convey any such inference, but on the contrary the capabilities of the apparatus and its limitations have been expressed in a perfectly definite way and should be easily understood by any experienced radio man. These may be summarized by stating that with this apparatus practically continuous reception from such stations as Carnarvon, Nauen, or Lyons is possible save only when there is local lightning present, while with any other known receiving method reception from these stations in the afternoon and evening during the summer is nearly always entirely impossible, and is also impossible to a considerable extent, at other seasons of

the year. In this statement it is assumed that the receiving station is within 100 miles (160 km.) of New York City.

Since the time of delivery of the paper the greater part of another summer season has passed, during which various forms of the static arrangement have been in continuous service. Furthermore, the static conditions experienced this year have been much more severe than those encountered last year, yet the continuity of reception from the stations mentioned has been greater than that of last summer and we have been able to receive continuously even thru the worst fading periods, which occur in the afternoon between four and seven o'clock, without interruption by anything except local lightning. With regard to this latter cause of interruption, to date there have been only two days on which lightning has rendered reception impossible; in one case for a period of approximately three hours, and in the second case for approximately two hours. There have also been several days—perhaps four in all—when lightning has caused appreciable trouble but not complete interruption; that is to say, thru these periods messages in plain language, sent twice, could have been received completely, but code, sent words once, could not have been completely copied.

Dr. de Forest's suggestion that the opinion of unbiased and qualified experts on the working of this system is desirable before its efficacy can be completely accepted, is entirely reasonable, and it was for this reason that the offer to conduct tests for the benefit of a committee to be appointed by the Institute was made. The response to this offer indicated that this was perhaps a rather difficult thing to carry out since a considerable period of observation would be necessary in order to arrive at a correct conclusion. With this point of view the writer is unable to differ since he was unwilling to accept the results himself until the system had been tested thru an entire summer, twenty-four hours a day, with an operator constantly on watch.

The most conclusive proof of the correctness or otherwise of the claims on the working of this system will be had when the Marconi Company resumes commercial working with Europe, after its stations are returned by the United States Government.

In reply to Dr. de Forest's statement that certain individuals in the Government service have informed him that the apparatus does not work as claimed, it is enough to state that only one man in the Government service has had any experience or fa-

miliarity with the system, and that this gentleman, Mr. George H. Clark, has already set forth the result of his observations, in his discussion of my paper.

With regard to the criticism that the arrangements used are of "brobdignagian" dimensions, it would seem pertinent to inquire how long it has been such a colossal task to construct six miles (10 km.) or so of ordinary telephone line, which, from the constructional point of view, is all the largest arrangement requires. Dr. de Forest has often boasted of the part his audion has played in transcontinental telephony, which involves some 3,000 miles (4,800 km.) of this same sort of construction. The cost is certainly not prohibitive, for the total expense of the Lakewood installation was less than one-half of that of a single tower of the type which the Marconi Company had at its Belmar receiving station, six of which were considered necessary for reception a few years ago.

Dr. de Forest has referred to some of the later arrangements as being far simpler and more rational because of their smaller size, but while this is true the fact is that none of them has yet been developed to the point where it is equal to the Lakewood arrangement in the range of conditions it is able to cope with.

With reference to this historical references of the paper, it is sufficient to say that, as specifically stated, no attempt was made to present a complete historical outline. There can be no ground for charging unfairness, therefore, and it is not necessary here to either accept, deny, or qualify Dr. de Forest's statements.

The writer takes issue squarely with Dr. de Forest in his statement that the Fessenden interference preventer circuit "where enlightened methods are employed" causes a great improvement in signal-static ratio, or that it is possible to secure with this arrangement any signal-static ratio which cannot be duplicated with other well-known receiving methods used with equal enlightenment.

He is also entirely incorrect when he states that that principle is relied upon to achieve the final elimination of static impulses between the static tank and the static signal circuits, for the reason that in this latter, one circuit has both signal and static currents, while the other has only static currents; consequently when the two are connected together in opposition, signal current only is left. If both circuits had both signal and static currents, then his statement would be correct.

Dr. de Forest is quite right when he states that if the hypo-

thesis of overhead origin and vertical propagation were correct it should be possible to eliminate static by rotating a small loop-receiving antenna about a horizontal axis. In fact this does accomplish the elimination of static, but unfortunately it also accomplishes the elimination of signal. If it were possible to transmit waves which were horizontally polarized instead of vertically, as at present, and if they would remain so at great distances, then a loop with its plane horizontal would furnish an ideal means of working thru static of the grinders type.

Dr. de Forest also refers to certain tests recently made by Major Charles A. Culver, in which he shows that static currents produced by two loops at right angles will not balance. If he had read the paper more closely he would have found this identical fact stated (at bottom of page 216 and top of page 217 of the PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, June, 1919). It was discovered by the writer more than three years before the work which Major Culver reports was undertaken. His further statement that if the hypothesis of vertical propagation were correct, loops at right angles should balance, is also entirely incorrect, as will appear from a less superficial analysis of the proposition. If it be assumed that these electromagnetic waves are generated by linear oscillators, then, in order that two loops at right angles to each other shall be affected by the wave originating at each oscillator simultaneously and with the same relative intensity, which is the prerequisite to balancing, the horizontal projection of each of these oscillators must make the same angle with the two loops, since if different impulses were due to oscillators, the horizontal projection of whose axes may be at different angles with the two loops, there would be no possibility of securing adjustment, which would annul any appreciable percentage of them. The conception that these oscillators are so arranged seems to me utterly impossible, and in order that the hypothesis of overhead origin and vertical propagation may be rational it is necessary to assume that the axes of the oscillators producing static waves shall assume all possible angles in space; that is, they shall be heterogeneous in their disposition. Under this assumption it becomes impossible for loops in different planes to balance. At this point it should be emphasized, as it was in the paper, that this hypothesis is set up merely as a convenient working base, and that it cannot be regarded as proven. On the other hand, however, the fact that antennas situated considerable distances apart are sim-

ultaneously affected by the type of static known as grinders, is proven beyond question.

Dr. de Forest's reference to the work of Latour in receiving with a small closed loop, and his comparison with the reception of Carnarvon's signal thru the interference of New Brunswick, is quite beside the point for the simple reason that the arrangement of Latour is in no sense a static eliminator, but is merely a small loop-receiving arrangement, whereas it was definitely stated in the paper that the result mentioned was accomplished *while maintaining a good static balance*, which is quite a different proposition.

His reference to Commander Taylor's results obtained with loops only three meters (9 feet) square also has no bearing since the arrangement described by Lieutenant-Commander Taylor is not a static preventer but a simple loop-receiving arrangement.

With regard to the cosine and other curves which Dr. de Forest scoffs at, it is suggested that he study them a little more closely until he discovers their significance, as his comment thereon indicates so complete a lack of understanding that reply is useless. If he so entirely fails to understand the paper he would equally fail to comprehend the reply.

His lengthy statements relative to the necessity for shielding the receiving apparatus itself from the direct effect of static are plausible but unsound. While it is a very common observation to note that considerable static and even signal is heard at a receiving station when the antenna switches are open, most of this is due to the electro-static coupling which exists between the aerial and the receiving instruments, and that part of it which is due to direct action on the receiving coils themselves proves to be so small as to be entirely negligible, provided an aerial of relatively large dimensions is used. If the dimensions of the aerial are reduced to the extreme limit, then of course the capabilities of the receiving coils become comparable to the capabilities of the antenna for picking up both signal and static, but extensive experience with this has shown that even with a loop aerial only four feet square the order of this effect is not sufficient to be serious, whereas with large aerials such as the Lakewood installation it is totally negligible. In the working of this latter system it is at times an interesting observation to note that the amount of static disturbances which is heard after a proper balance is obtained, is less than that which is heard when all three aerials are disconnected from the receiving set.

In conclusion, Dr. de Forest again intimates that there are

still certain to occur periods when the only way in which we can get rid of static is to lay the telephones on the table. No such periods, with the exception of those caused by local lightning, have occurred during the prolonged use to which this apparatus has been subjected, and it therefore does not yet appear that Dr. de Forest is correct. The offer to demonstrate made in the paper still stands, and Dr. de Forest himself is not excluded.

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PROCEEDINGS

of

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TECHNICAL PAPERS AND DISCUSSIONS

EDITED BY
ALFRED N. GOLDSMITH, Ph.D.

PUBLISHED EVERY TWO MONTHS BY
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Corrections: On the following pages of the 1919 PROCEEDINGS, page 453, line 1 should read:

"decreases from *d* toward the point *n*"

Page 478, line 12 from bottom, change "affect" to "effect"

Page 488, lines 14 and 15 should read:

"because of the rate at which"

At the end of this number are the title page, page of general information, and table of contents pages for the entire Volume 7 (1919) of the PROCEEDINGS. These last may be suitably placed at the beginning of the volume for binding.

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LONG WAVE RECEPTION AND THE ELIMINATION OF STRAYS ON GROUND WIRES (SUBTERRANEAN AND SUBMARINE)*

By

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Many of the properties of ground wires with respect to long wave reception have been touched upon in the previous paper which dealt mainly with short wave work (Proceedings of The Institute of Radio Engineers, volume 7, number 4, 1919). The purpose of this paper is to take up some of the special problems of long wave reception with ground wires, with special reference to the work done by the writer on the elimination of strays.

1. Optimum Wire Length for Long Waves

Number 12 rubber covered wire[1] was used for all of the earlier experiments on optimum wire length because it was found to hold its insulation for several weeks and was cheap and easy to handle. It is not recommended for permanent installations. It has already been shown in the previous paper that for 600 meters the optimum length for this wire was 125 feet (38.1 meters) each way, and that up to 1,125 meters this length seemed to be proportional to the wave length. It was therefore expected that a similar relation would hold for waves between 4,000 and 15,000 meters. For 12,000 meters the length was therefore expected to be 2,500 feet (763 meters). Since there was comparatively little arc work being done by stations south of Great Lakes, and since there were no arc stations north of Great Lakes, it was necessary in the work done there, to attempt optimum length experiments on stations either east or west of Great Lakes. At the laboratory on the bluff, it was not possible to lay wires in trenches for so great a distance, while at the station on the beach, it was only possible to lay a wire in one direction, using

* Received by the Editor, March 7, 1919.

[1] Diameter of number 12 wire = 0.081 inch = 0.205 cm.

it against a ground. An attempt was made in two ways to determine whether or not optimum length existed for these long waves. First the signals from Lyons, France, on 15,000 meters were observed on a wire 3,000 feet (915 m.) long, running straight east into the lake, the outer end of the wire being sixty feet (18.3) under water. This wire was gradually pulled in and observations taken. It was a laborious and difficult matter to obtain satisfactory observations in this way, but those that were taken indicated that 2,650 feet (808 m.) gave the best signal for Lyons. The signals were too weak to get, with the amplification at that time available, any adequate measure which would indicate whether the ratio of signal to stray was better at this length than at others. About this time Doctor L. W. Austin reported that, as far as he could determine from the experiments made in the slightly brackish water of the Potomac at Anacostia, District of Columbia, there was no optimum wire length for long waves and that no proportionate increase in signal was observed after 2,000 feet (610 m.). In the Great Lakes experiments, all signals were compared with those received on a standard wire, 2,000 feet (610 m.) in length. In order to avoid the laborious process of hauling in the long wire, which occupied considerable time, the problem was attacked at Great Lakes on a different basis. Two wires, separated 50 or 60 feet (15 or 18 m.), running in the same direction were compared. They were both fixed in length, one being 2,000 feet (610 m.) and the other 1,750 feet (534 m.) long. For various wave lengths between 5,000 and 14,000 meters, the ratio of signals on the 2,000 foot (610 m.) wire to signals on the 1,750 foot (534 m.) wire was determined. These observations were insufficient in number to be at all conclusive, but the best ratio was obtained at 12,600 meters, Nauen's wave, indicating that 2,000 feet (610 m.) was not far from the optimum length for this wave. It is, of course, possible that the relation between optimum wire length and wave length is not exactly linear, and it is deemed that the data herein reported is not entirely satisfactory. The experiments on optimum length were continued later at the U. S. Naval Radio Station, Belmar, New Jersey, which was then the principal station and control center of the trans-Atlantic system and where the writer was stationed as trans-Atlantic Communication Officer. The Belmar experiments on wires laid in the inlet (salt water) in front of the radio station, showed that up to the length of 1,500 feet (458 m.) signals from Nauen on 12,600 meters continued to increase. It was impossible to obtain a greater

distance than 1,500 feet (458 m.) without deviating too far from the proper direction. During the month of January, ice formed on the inlet and a piece of "packard cable," number 14 high tension[2], was laid on the surface of the ice for the purpose of determining the optimum length of Nauen's short wave, 6,300 meters. The signal strength rose rather rapidly until a thousand feet (305 m.) were used, after which it rose very slowly so that it was difficult to determine exactly where the optimum length lay. It was estimated to be 1,600 feet (488 m.). Similar experiments with a wire on the ice, using Lyons' spark wave of 5,300 meters, indicated an optimum length of 1,200 feet (366 m.) and showed also that the rise of signal strength was very gradual and that there was no practical advantage in using over 800 feet (344 m.) of wire for 5,000 meters and not over 1,000 feet (305 m.) for 6,000 meters. About this same time, January, 1918, lead covered cable on the surface of the ground was tested at Belmar. The sheath of the cable was grounded at a number of points, special care being taken to get a good ground at the receiving end. The core of the cable contained two number 18 copper wires[3], which were connected to the receiving set and used against a ground connection. The behavior of lead-covered cable showed at once that the most suitable length for long waves was decidedly different from that proper for ground wires or submerged wires. For instance, while 2,000 feet (610 m.) of underground wire was found very suitable for waves of 10,000 meters and upwards, it was found that a lead-covered cable 3,000 feet (915 m.) in length showed up best on wave lengths between 5,000 and 6,000 meters. A lead-covered cable 7,000 feet (2,135 m.) in length was then opened at a series of points 500 feet (153 m.) apart and observations were taken on Nauen's 6,300 meter wave, comparison being made in each case with the signals obtained on a fixed 2,000-foot (610 m.) ground wire. A curve was plotted from this data which showed a maximum at 3,000 feet (915 m.); the curve was, however, very flat. A little later experiments were undertaken with a ground wire buried seven feet (2.1 m.) deep, a number of pits having been dug for the installation of disconnecting switches. Observations were taken on signals on 9,500 meters from Stavanger, Norway, and on Nauen's 6,300 meter wave. The total length of ground wire available was 2,000 feet (610 m.). The observations were inconclusive, the Stavanger signals at 9,500 meters indicating a

[2] Diameter of number 14 wire = 0.064 inch = 0.162 cm.
[3] Diameter of number 18 wire = 0.040 inch = 0.102 cm.

maximum when the full length of the wire was used, whereas measurements on Nauen indicated a linear rise proportionate to the length of the wire, that is to say, the Nauen observations indicated no optimum length inside of 2,000 feet (610 m.). It is regretted that the pressure of other work interrupted these interesting experiments at this point. A little later in the Spring, experiments were begun at the Naval Radio Station, Chatham, Massachusetts, by Gunner D. J. Burke, under the writer's direction, and some attempts were made there to discover whether there was such a thing as optimum wire length in fresh water, salt water, and in ground. The results were negative as far as salt water is concerned, and doubtful as far as fresh water and ground were concerned. During the summer a special station was erected at Belmar with a 2,000-foot (610 m.) sea wire, but no results were obtained with this wire indicating an optimum length. Altho an insufficient number of observations on long waves have been taken, it is quite evident that the optimum length does not exist in salt or brackish water and that if it exists in fresh water or for wires buried in the ground, the optimum is not at all sharp. The most positive indications of optimum length on long waves were obtained with surface cables laid either on the ground or on the ice. Theoretical considerations would indicate that for very long waves and correspondingly long wires, the resistance of the system is so high that it is much more perfectly aperiodic than is the case for short waves. Confirming this is the fact that a single ground wire may be so readily used for the reception of a number of different stations on different wave lengths without any interference whatever in the tuning, it being assumed, of course, that the local oscillations of the receiving bulbs are so adjusted as not to heterodyne against each other. It was frequently possible at Belmar to copy simultaneously on any of the ground wires or sea wires, Lyons on 15,000 meters, Carnarvon on 14,000 meters, Nauen on 12,600 meters, Rome on 11,000 meters and Nantes on 10,000 meters. This procedure, altho possible on short waves, does not work out well at all, because each wave requires, for best results, its particular optimum length.

2. Ratio of Signals to Strays

In order for the ground wire system to be of practical value it must be able to show advantage in readability of signals not only over an ordinary aerial but over a properly designed receiving frame or closed loop, since the latter is more compact

and easier of installation. The elimination of *actual static* is, of course, fairly complete on the ground wires, but the relative advantage of ground wires over rectangles, as far as the elimination of *all* strays was concerned, had to be made the subject of exhaustive tests. Early in January, 1918, the writer requested Ensign A. Crossley at Great Lakes to construct a rectangle 11 feet (3.36 m.) square, wound with 80 turns of number 13 double cotton-covered wire[4] spaced 0.5 inch (1.27 cm.) apart. This rectangle was compared for a considerable period of time with the 1,200 foot (366 m.) "packard cable" at the Great Lakes laboratory station on the bluff, the cable being buried four feet (1.22 m.) under the surface of the earth. The ground wires gave signals averaging three times as strong as those on the rectangle. The ground at that time was partly frozen. The following table is typical of the observations obtained at Great Lakes:

Station		Ground Wire	Rect-angle	Ground / Rect.
WGG	Tuckerton, New Jersey	5.5	4.7	1.17
NPL	San Diego, California	6.6	2.9	2.26
KET	Bolinas, California	2.2	1.3	1.70
NAA	Arlington, Virginia	1.5	2.8	0.54
NAD	Boston, Massachusetts	1.0	1.4	0.72
KIE	Heiia Point, Hawaii	0.4	0.3	1.33
NBA	Darien, Panama Canal Zone	1.3	0.9	1.45
NPC	Puget Sound, Washington	2.2	1.6	1.38
KSS	San Francisco, California	2.1	1.2	1.75
NPM	Pearl Harbor, Hawaii	1.5	0.7	2.13
NPG	San Francisco, California	1.0	0.7	1.44
WII	New Brunswick, New Jersey	15.0	7.4	2.02
POZ	Nauen, Germany	0.2	0.1	2.00
BZZ	Carnarvon, Wales	0.3	0.2	1.50
NPA	Cordova, Alaska	0.3	0.1	3.00
			Average—1.626	

The average readability of signals at Great Lakes was 62.6 per cent. better on the ground wire than on the rectangle. About the time the frost penetrated well into the ground at Great Lakes, it had been noted that the strays became distinctly worse. The same thing was noticed on the sea wires at Belmar when the

[4] Diameter of number 13 wire = 0.072 inch = 0.183 cm.

shallow inlet froze up so that the wires were partly covered with a three-inch sheet of ice. In order to get further evidence, wires were laid at Belmar on top of the ice and directly over the sea wires and the ratios of signals to strays on many trans-Atlantic stations were obtained in comparison with the signals on the sea wires frozen in the ice. The readability of signals, defining readability as the ratio of signals to strays, was twice as good on the sea wires under the ice, altho not as good as on the same wires without any ice over them. In the meantime hundreds of observations had been accumulated at Belmar comparing the ratio of signals to strays received on rectangles 77 feet (23.5 m.) long by 30 feet (9.2 m.) high, with 12 turns of number 10 copper wire[5] spaced 6 inches (15.2 cm.) apart, with those obtained on 1,200, 1,400, and 1,700-foot (366, 427, and 519 m.) sea wires and with those obtained on a 2,000-foot (610 m.) land wires buried 2 feet (61 cm.) deep. Many observations were also made on a 2,000-foot (610 m.) land wire buried 7 feet (2.14 m.) deep. This latter wire gave louder signals than the one buried 2 feet (61 cm.), but the same ratio of signals to strays. The general average showed that the signals obtained on rectangles and sea wires were of approximately the same intensity, but that the readability of the signals received on the sea wires was twice that received on the rectangles. On the other hand, the ground wires, altho giving signals four to five times as strong as the rectangle, showed no advantage whatever in readability. Similar experiments were carried out during the summer at the Naval Radio Station at Tuckerton, New Jersey, with the ground wires placed in extremely moist earth and where they showed a marked advantage over the rectangle, altho they proved to be not quite as good as the sea wires at Belmar. In the meantime a great many measurements had been made at Chatham, Massachusetts, on wires both in fresh and salt water. The fresh water wires showed tremendous signals but no better ratios or readability than rectangles. The sea wires on the other hand showed good readability, but owing to their being covered part of the time by a high tide to a depth of six feet (1.8 m.), they showed rather weak signals. It was attempted to remedy this by suspending them from floats, but owing to interference with traffic in the bay and to stormy weather conditions this was abandoned as being impracticable. It must be noted that the good results on long waves in the earlier experiments at Great Lakes were obtained with wires buried in wet sand and a little later with wires buried on the bluff at a

[5] Diameter of number 10 wire = 0.102 inch = 0.259 cm.

sufficient depth to be near ground water level, in fact below it at a good many points. It is quite evident that the ground wire system possesses no advantage over the rectangle in the elimination of strays other than static, except when the ground wires are laid in a partially conducting medium. An attempt was made to confirm these results at the Naval Radio Station at Bar Harbor, Maine, and it was found that ground wires laid in the very rocky surface soil seemed to have even worse strays than those received on the rectangle. A sea wire, 1,100 feet (1,336 m.) long was placed in an inlet, but was found to be, even when floated on the surface, completely shielded from all trans-Atlantic signals, there being a cliff considerably over 100 feet (30.5 m.) high on one side and another cliff 80 feet (24.4 m.) high on the other. The waves apparently jumped this gap without influencing the wire floating on the surface of the water in the least, showing a rather interesting case of complete shielding. The effect of the freezing of the ground or of the water in which the wires were placed is evidently due to the change in conductivity thereby produced. Bearing these facts in mind an attempt was made at Belmar to use lead-covered cable in such a way as to imitate the properties of properly installed ground wires, without the necessity of burying them. These attempts met with a partial degree of success. A lead-covered cable showed stray ratios intermediate between those of the wires buried in dry soil and the sea wires. If, however, the sheath of the cable was frequently intercepted so that its electrical continuity was broken up, the signal rose greatly in intensity and the strays still more so, until the ratio of signals to strays was slightly worse than that obtainable upon wires buried in dry soil. If the conducting medium surrounding the wires is of too high a conductivity, good results will not be obtained, especially if the wire is lowered to any considerable distance below the surface. During the month of June, 1918, experiments were made ten miles (16 km.) off the coast at Belmar with an 800-foot (244 m.) wire trailed behind a small motor dory. The signals from trans-Atlantic stations decreased rapidly with the depth of the wire, so that at 15 feet (4.6 m.) below the surface, it was not possible to copy them with two stages of amplification. Of course, trans-Atlantic signals at that time of the year and at mid-day, when the test was conducted, were not very strong, nevertheless when the wire was within four feet (1.2 m.) of the surface, Carnarvon, Nauen, Nantes, and Rome were all copied without difficulty on this particular occasion.

3. Methods of Eliminating Strays From Long Wave Receivers

It may be of interest to note here a few of the methods tried out at Belmar for the further suppression of strays on ground wires and particularly on sea wires, altho the principle was finally accepted that the strays had to be eliminated before entering the receiving circuit.

(a) Tuned telephone circuit. The use of group tuners or audio frequency tuners in the amplifier circuits was tried out and shown to have some slight advantage, but not enough to warrant its general adoption. The tendency of all such devices is to produce a ringing or blurred signal. These devices are very deceptive to the ear; they appear at first to produce a very great improvement, but when one tries to make copy it is discovered that the improvement is generally imaginary.

(b) Audio frequency balanced circuits. A device somewhat similar to the Fessenden interference preventer was next tried out. It consisted of two complete receiving circuits arranged in duplex thru a differential transformer which led to the amplifier. The idea here was to take advantage of the fact that several circuits could be tuned to one ground wire without mutual disturbances and by tuning one of the circuits, say to 14,000 meters and the other to 13,500 meters, and opposing the output of the receiving bulbs in the differential transformer, it might be possible to balance out strays on adjacent wave lengths without eliminating the signal. It was found, however, that this was not possible unless the signal as well as the strays was balanced out.

(c) Radio frequency balance. A radio frequency differential transformer was then tried out, the device consisting of two circuits tuned to nearby waves, the secondaries of the two tuners being coupled differentially to a tertiary circuit, the coupling being made very loose. Some improvement was obtained with this circuit, altho the adjustment was very critical in order to get exact opposition in phase and amplitude. It had one great advantage, namely, it was possible to differentiate very sharply between stations the wave lengths of which were very close together, but on the whole it was not considered to be of sufficient assistance in the elimination of strays to make it worth while.

(d) Radio frequency amplifiers. A three-stage radio frequency amplifier with tuned circuits and very loose coupling was next tried out. Again the results showed very high selectivity and a noticeable gain in readability, nevertheless the circuits

were so difficult to handle that it was not considered to be satisfactory from an operating point of view, neither was the gain sufficient to make it seem worth while.

(e) Automatic recorders.

An automatic recorder of unusually good design, selectivity, and sensitiveness was then tried out. At first it seemed to promise great results, but after reading several thousand feet of tape and comparing it with the copy obtained by a good operator, it was found that the recorder was no more reliable than a good operator and if the speed was increased beyond thirty words a minute, the recorder was badly interfered with by strays. One difficulty was that rather excessive amplification was necessary in order to use this recorder on sea wires.

(f) The use of high pitched telephones. Some experiments of considerable interest were carried out with very high pitched telephones and these experiments did show a bona fide improvement in the ratio of signals to strays and the writer believes that there is a profitable field of investigation in this line, if telephones of high pitch can be manufactured with a sensibility that is comparable with that of standard telephones at a thousand cycles.

4. The Elimination of Strays from Land and Sea Wires

It was finally decided that the only way to do anything towards the further suppression of strays over and above that already obtained by the use of a good sea wire, was to apply the method of elimination ahead of the primary of the receiver. Considerable improvement in the ratio of signal to stray was obtained by placing a low resistance of value between 1 and 25 ohms across the primary of the receiving set. It will be remembered that the receiving sets were standard Navy long wave tuners and, therefore, had a series condenser, and that in ground wire work the tuning of the primary is dependent only upon the constants of the primary and not upon the length of the ground wire, the only exception being when exceedingly short ground wires are used. There is also probably some slight deviation from the rule when working with short waves around the optimum wire length. The placing of the shunt around the primary therefore did not in any way affect the tuning of it. The improvement in signal-to-stray ratio obtained by the use of the shunt is at the cost of considerable diminution in signal strength and is not therefore of very great value in improvement in readability except in special cases. It having been determined that

the sea wires had twice as good a ratio of signal to stray as the rectangles, it seemed likely that one ought to be able to balance the strays from a rectangle against those from a sea wire and still have some signal left over. This was first attempted by coupling magnetically the primary of the receiving set by means of a differential radio frequency transformer to both sea wire and rectangle. The differential transformer had one secondary coil which was in series with the primary of a receiving set and it had two primary coils, one of which by means of a series condenser was tuned to a rectangle and the other tuned by means of another series condenser and suitable loading coils to one of the sea wires. See Figure 1. As a balancing arrangement the device worked perfectly, but altho signals even of very great intensity could be accurately balanced out, it was not possible to balance out strays. It should be noted that the planes of

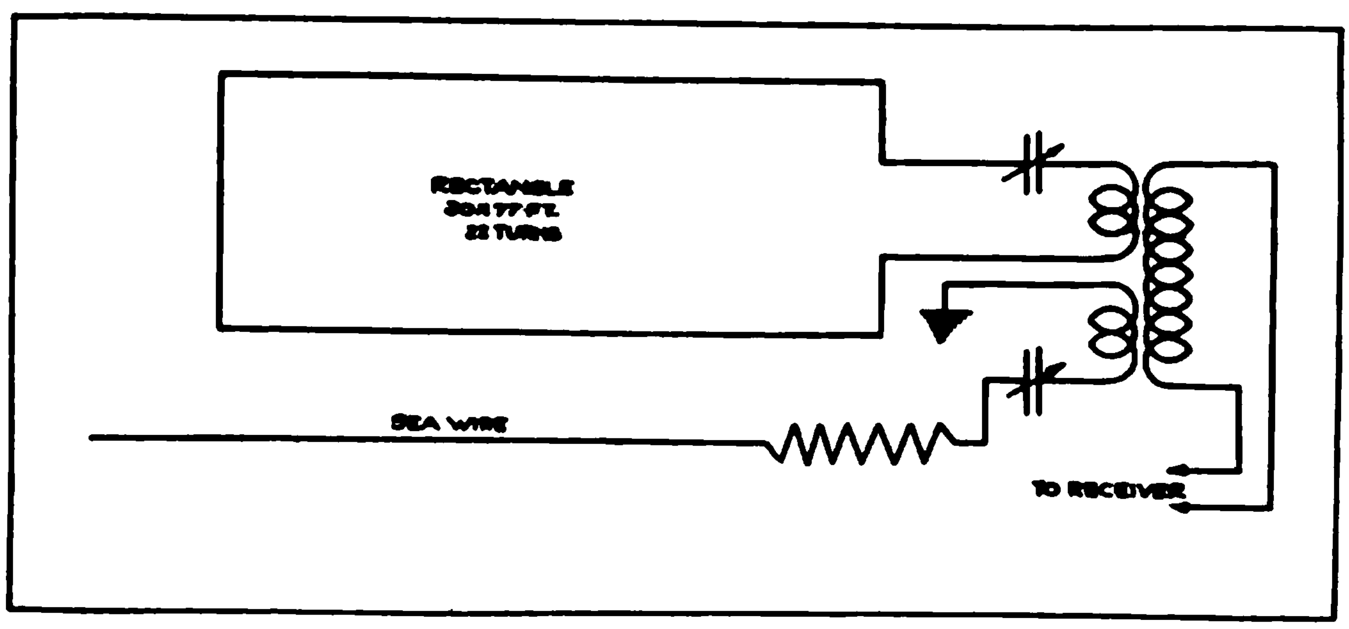


FIGURE 1

the rectangle pointed in the same direction as the sea wire, that is towards the European stations. The failure of the experiment was, at the time, laid to a lack of exact similarity in directive properties of the two component parts of the balanced system, but it is the present opinion of the writer that the failure was due to the fact that the rectangle constitutes a relatively feebly damped receiving system, while the sea wire is, especially for long waves, aperiodic. A similar attempt, shown in Figure 2, was made to balance a land wire against a sea wire and with the same results as far as this circuit is concerned. This is probably due to the fact that a land wire in dry soil is not so nearly aperiodic as a sea wire. If the land wire were laid in wet soil the experiment would also fail, because the ratio of

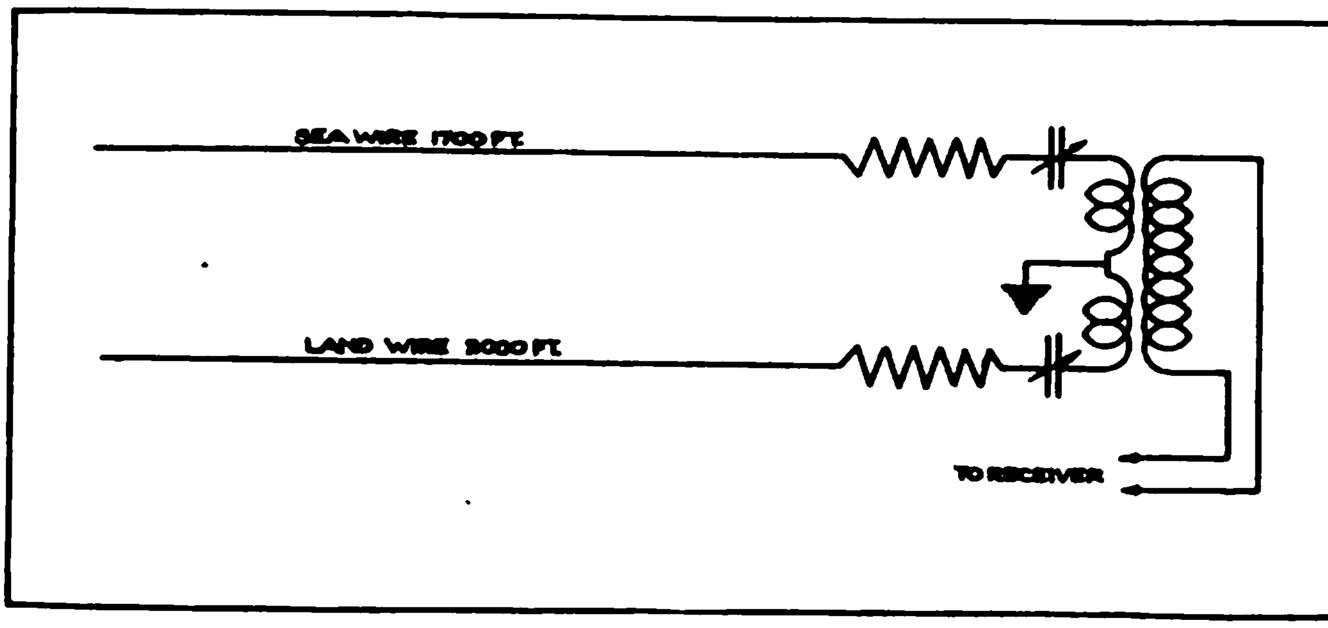


FIGURE 2

signal to stray would be too nearly the same for both sides of the system. The next attempt, shown in Figure 3, was to balance by means of a small potentiometer arrangement, a land wire against a sea wire. The resistance *R* was a slide wire rheostat, various values being tried from 50 to 2,000 ohms. The

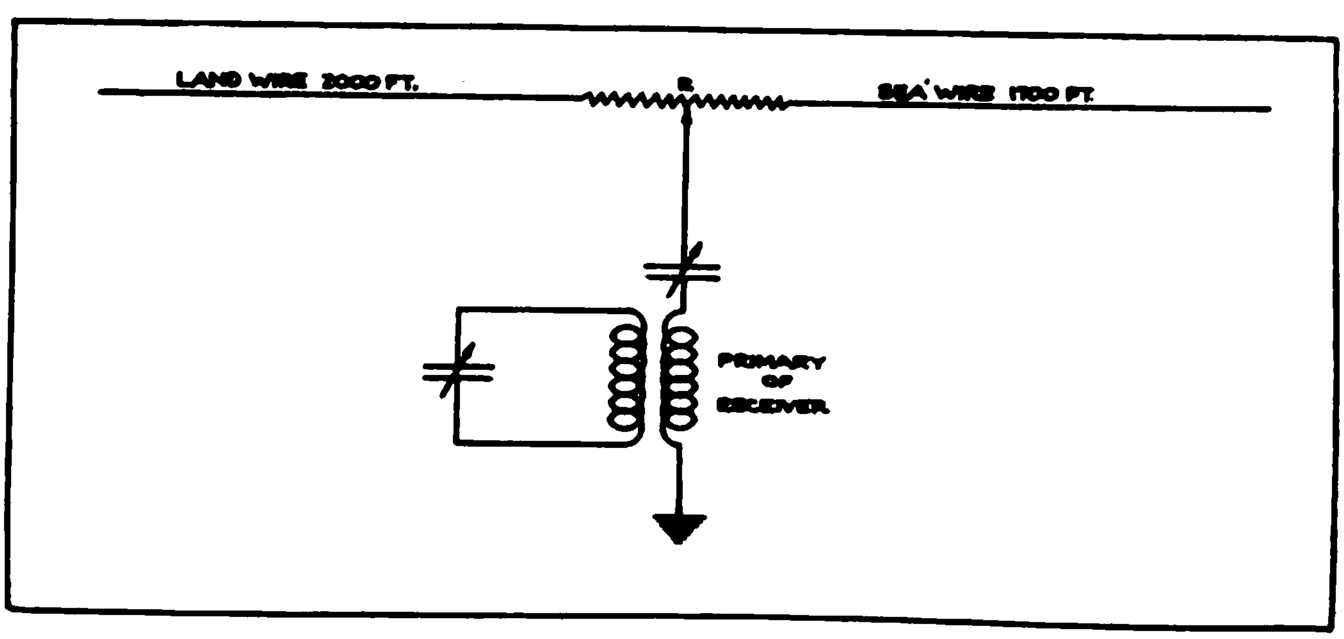


FIGURE 3

idea in the arrangement of Figure 3 was that the current from sea wire to ground would be opposed in phase from the current from land wire to ground and by suitably proportioning the two parts of the resistance *R*, the strays could be balanced out. Such, however, did not prove to be the case and it was recognized that the difficulty was due to the fact that the phase relationship of the current in the sea wire was not the same as for the current in the land wire. The final arrangement for the balance of the land wire against the sea wire is shown in Figure 4, where a phase-adjusting device, $L_1 C_1$, is put in series with either land wire or sea wire. It is not necessary to have this

device in series with each collector. It was usually used in series with the land wire. It will be noted that the ratio of signal to stray on collector number 1, that is, the sea wire, has been further improved by the use of the resistance R_1, which shunts the end of that wire directly to ground. This circuit at once

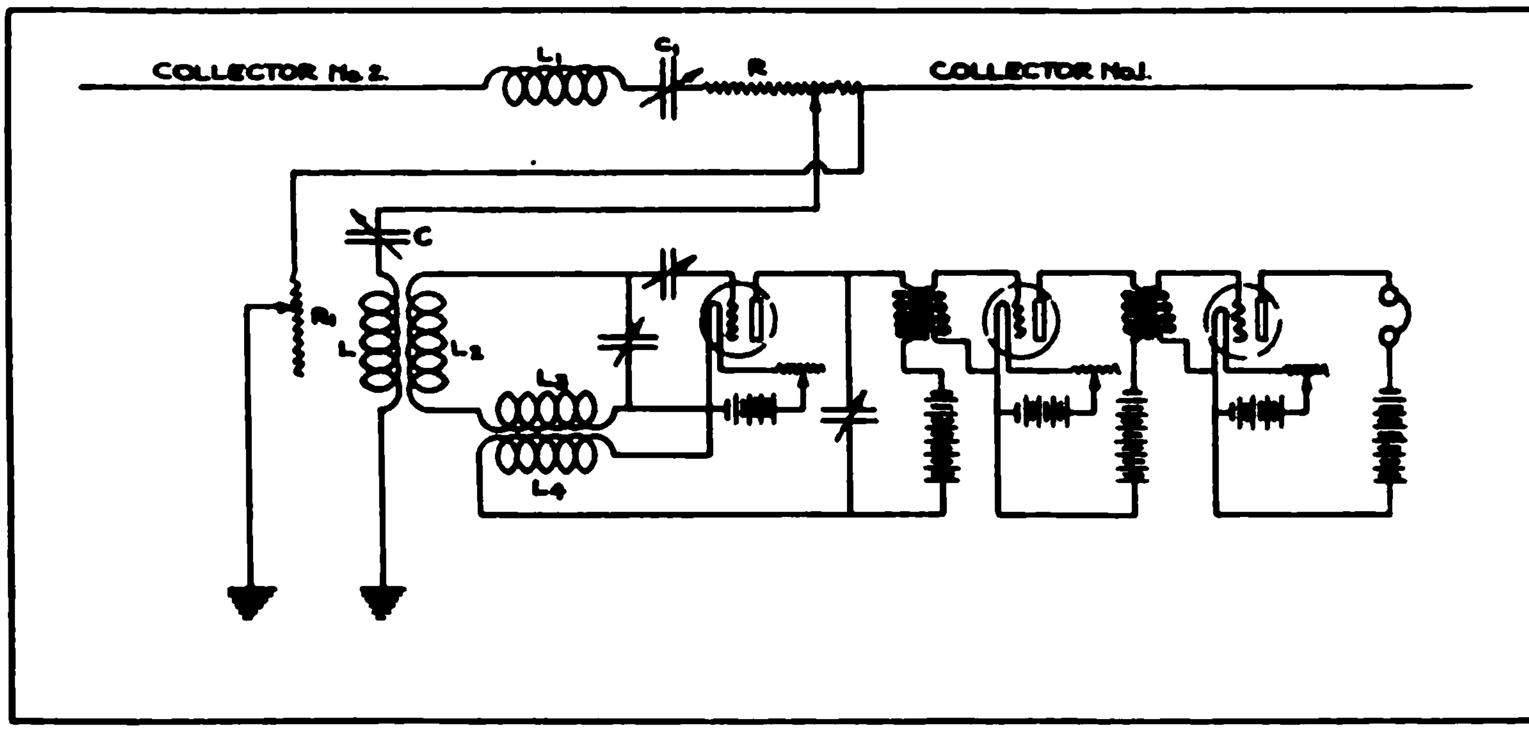


FIGURE 4

gave very satisfactory results and experiments were immediately continued to determine whether the balance of strays was dependent upon the length and nature of the land and sea wires. It was found that within the limits wherein observations were taken on sea wires, namely from 1,000 to 2,000 feet (305 to 610 m.), the length of the sea wire had but little influence, best results being obtained with 1,500 feet (458 m.) of sea wire. The length of the land wire was found to depend upon the wave length, in general, shorter lengths working better when shorter waves were to be balanced. The lead-covered cable was used as a land wire very successfully when the strays were made as bad as possible on the cable by intersecting the sheath, every two hundred feet (61 m.). It is desirable, of course, that the land wire have as bad a ratio as possible. Exceedingly satisfactory balances were obtained thruout the summer of 1918 on all wave length between 6,000 and 15,000 meters. No work was done on shorter wave lengths than 6,000 meters. The device was put into the hands of the operators on April 7, 1918, and either this circuit or the following circuit was used at Belmar from that time on for copying all trans-Atlantic signals. Encouraged by the success of this method of balancing two wires, attention was again given to the rectangle, since the rectangle showed just as

bad a ratio of signal to stray as the land wire and should have sufficiently similar directivity. The rectangle was therefore substituted for the land wire. The resistance R is sufficiently high to give the rectangle a very high decrement. The tuning is exceedingly flat. No marked success was obtained until the circuit shown in Figure 5 was adopted, that is, until the sea wire was shunted to earth thru a small resistance R_1. The exact

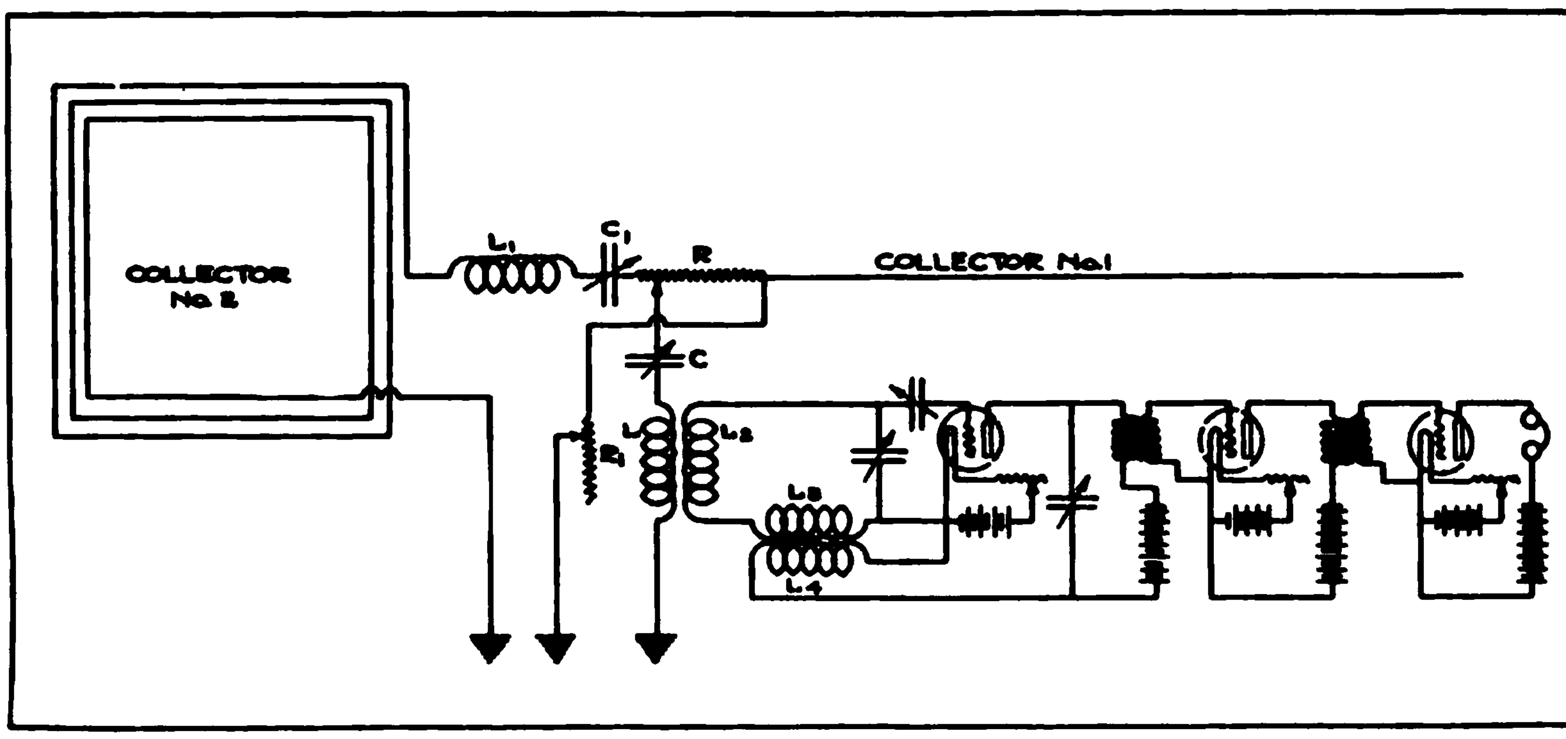


FIGURE 5

functioning of this resistance is not understood, but its presence, especially in the balancing of a rectangle against a sea wire, is of the utmost importance. Figure 6 shows the design of a panel to be placed to the left of a standard Navy long wave receiver, this panel providing the necessary terminals for sea wire and either land wire or rectangle. It also contains the phase-adjusting device in series with the rectangle or land wire, the balance

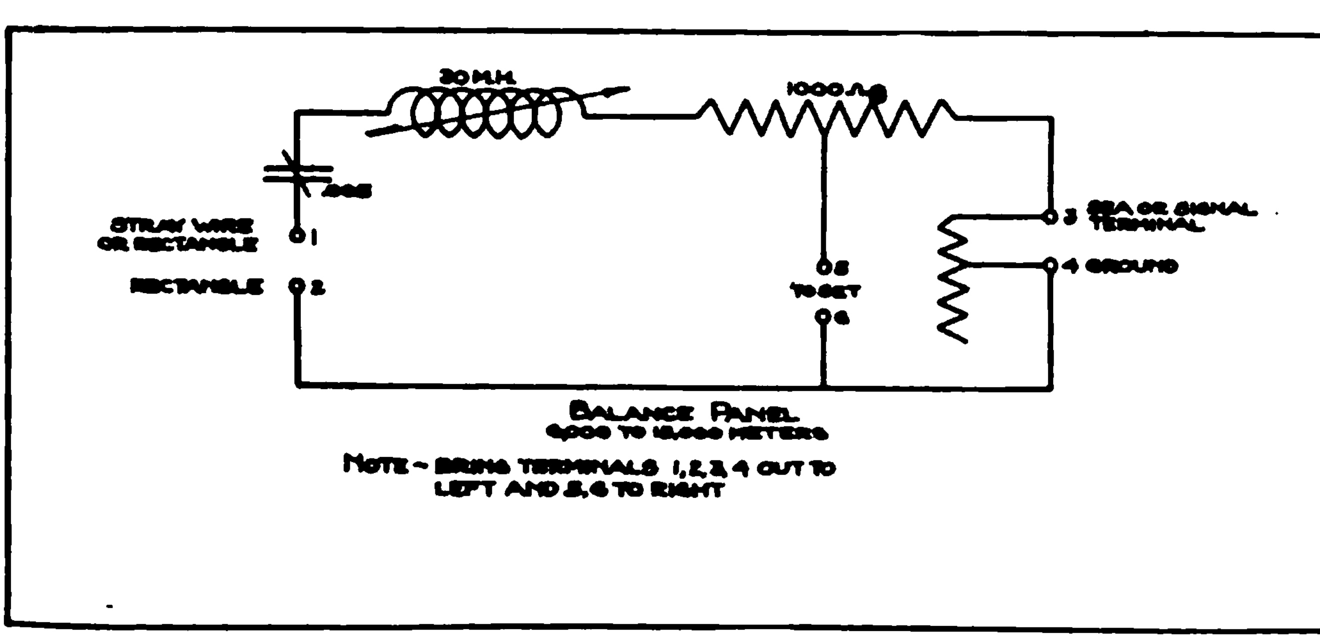


FIGURE 6

resistance, and the shunt to earth on the sea wire terminal. This circuit, Figure 5, is dependent also for its success upon the proper choice of the dimensions of the rectangle, since it is necessary to have a comparatively high resistance in series with it at all times. The rectangle must, therefore, be designed to have adequate collecting power. Naturally this depends upon the wave length. For 6,000 meters a rectangle 30 feet (9.2 m.) by 77 feet (23.5 m.) with 12 turns of number 10 wire[6] spaced 6 inches (15.2 cm.), was found satisfactory. For waves from 10,000 to 15,000 meters, a rectangle of the same dimensions but with double the number of turns was found best. The setting of the phase-adjusting condenser depends slightly on the depth of the water over a sea wire. As the tide came in it was found necessary to advance the phase slightly in the ground wire or rectangle, as the case might be.

5. Methods of Adjusting Balanced Systems

The method of adjusting the balanced system is described as follows (reference Figures 4 or 5):

(a) The slider of the resistance R is pushed to the right until there is little or no resistance between the slider and collector number 1. The primary of the receiver is adjusted by variation of the inductance L and the capacity C until it is tuned to the incoming signal. The resistance R_1 is adjusted to the lowest value which is consistent with good audibility of signal. The secondary L_2 is adjusted in the usual manner as are also the amplifiers.

(b) The slider of the resistance R is pushed to the left so that little or no resistance lies between the capacity C and the primary. Without changing the primary adjustment, the loading coil L_1 and the capacity C_1 are adjusted so that the same signal is received from collector number 2.

(c) The slider of the resistance R is then moved back and forth until the best readability of signals is obtained, the normal position being nearer the end of collector number 1 than to the capacity C_1. In other words, the larger part of the resistance R will normally be in series with that collector which produces the worst strays.

(d) The condenser C_1 is now varied so as to shift the phase slightly in collector number 2. This adjustment is fairly broad, as on account of the resistance R, the tuning in the circuit involving collector number 2 is extremely broad, in fact the cir-

[6] Diameter of number 10 wire = 0.102 inch = 0.259 cm.

cuit is almost, if not quite, aperiodic. After a few adjustments have been made, the balance of the circuit is extremely simple, reminding one forcibly of the method of balancing employed in the bridge method for comparison of inductance at audio frequencies, where variable resistances and one variable inductance are used and the other unknown inductance is fixed. The difficulty of balance is of exactly the same order, which means that after a very little experience, it is not difficult at all. In fact, after one or two days' training, this system was put into the hands of operators who had had comparatively little experience, men who had been thru the Naval Radio School at Harvard University and whose only practical experience was that which had been acquired in the course of duty at the Belmar station. From time to time slight corrections in the balance may be advisable, as the character of the strays changes. These corrections are, however, mostly in the phase-adjusting condenser C_1, and were thought to be due largely to the influence of the tide in shifting the phase of the signal in the sea wire.

6. Character of Strays

The behavior of the balanced system is such as to lead to the conclusion that strays are very complex. Certain very sharp and violent strays were soon recognized, after experience with this set, to be of comparatively local origin, traceable to some storm within a radius of about one hundred miles (160 km.). It would frequently happen that it was possible to obtain trans-Atlantic copy when there were violent storms in the immediate vicinity of the station. The lightning flashes themselves would produce very brief and sharp crashes which also manifested themselves at times by discharges thru the lightning arrestors in series with ground wires. But these brief disturbances constituted only a small interruption in traffic. At other times, these local storms which could be seen or heard from the station, produced quite complete interruption. It was finally discovered that if the storm was approaching from a direction at right angles to the system, it produced far less interruption to traffic than when it approached parallel to the system. In other words, the system has a certain amount of what may b " " "
cussing effect," the term being one to which I s
Mr. E. F. W. Alexanderson. Referring to Figure
that if the two parts of the resistance R are so ch
ance a disturbance ten miles (16 km.) away, tha
a distance comparable with the linear extent of

will not balance a disturbance a hundred miles (160 km.) away. Since the energy of the impulse varies as the inverse square of the distance from the centers of the respective collectors, number 1 and number 2, it must be evident that the adjustment of the slider of the resistance R, which equalizes impulses from a nearby point, will be very different from the adjustment which equalizes disturbances from a great distance. The greater the length of the two collectors, number 1 and number 2, the more pronounced will be this focussing effect. It can be utilized to great advantage in distant control work, where the interfering station is within a few hundred feet (about a hundred meters), but the balance obtained for eliminating such local interference is not the same as that which is utilized for the elimination of strays. All of the evidence collected during the summer of 1918 tends to show that the origin of the strays, while varying widely, is, except for those produced by local storms, at a very considerable distance, probably several hundred miles (over about 500 km.). Rarely did it happen that strays were bad at Belmar but what they were reported bad also at Sayville and Chatham. Bar Harbor, however, had entirely different receiving conditions, and on the whole was so remarkably free from strays as to be on an entirely different basis. The balanced system has, in view of the aforesaid focussing properties, a great value in handling local disturbances and nearby interferences, but this advantage cannot be utilized to its fullest extent at the same time that the stray balancing property is made use of. In the event of the necessity of simultaneously coping with bad strays and nearby interference, a compromise balance has to be effected.

7. Ratio of Improvement

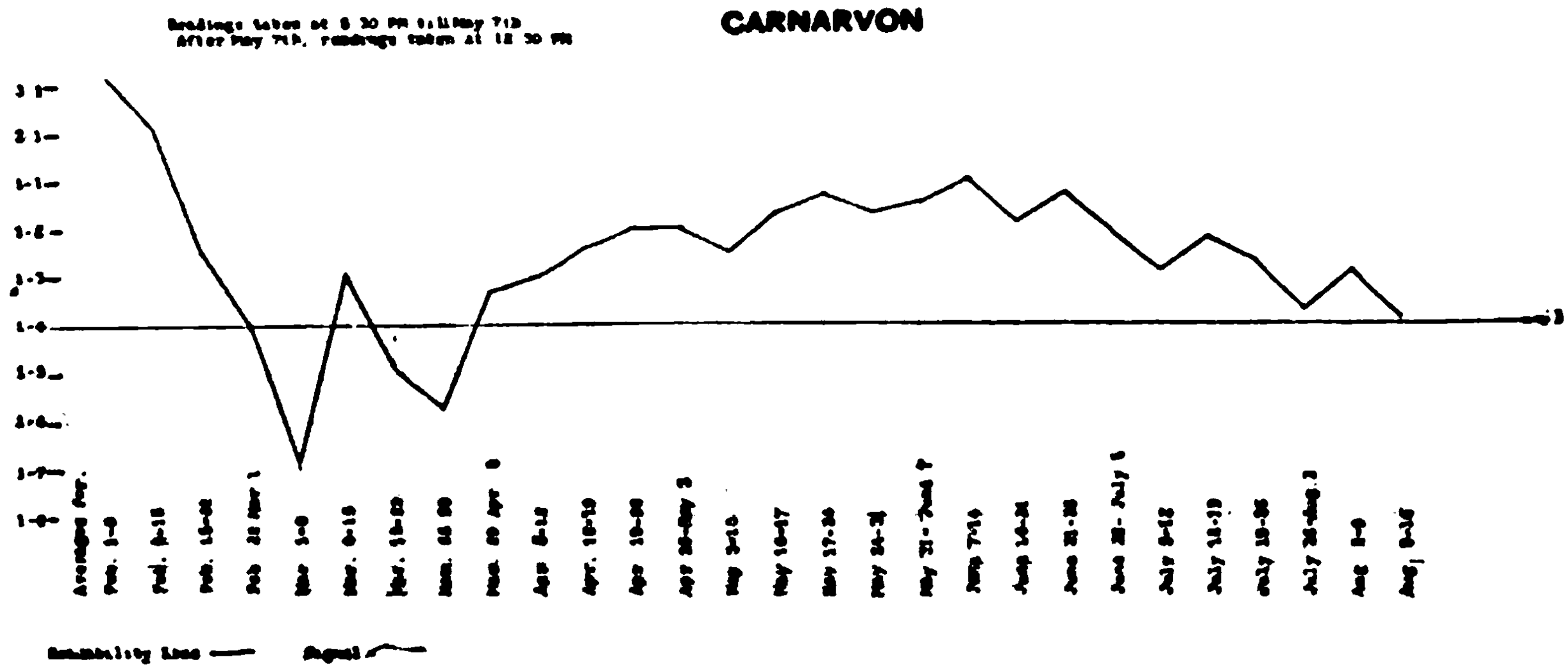
Multiplex reception of any number of stations on the same ground or sea wires is readily possible when using unbalanced reception, but with a balanced circuit it is necessary that each station be received on an independent pair of collectors. Six complete balanced circuits, some of the type shown in Figure 4 and some of the type shown in Figure 5, were therefore installed at Belmar and by the middle of April, 1918, all trans-Atlantic copy was made using the balanced system. Many observations were taken, showing the very great improvement in the readability of signals. These observations, in case of strays which are very heavy, are difficult to obtain. It has frequently been possible to get 95 per cent. copy using the balanced system on

signals which, unbalanced, were not even audible on account of absolutely continuous strays. Strays which arrive in an almost continuous stream are handled most effectively by the balanced system. The following table, reported to the Bureau of Steam Engineering under date of July 27, 1918, is typical of the results obtained. Since hundreds of observations have shown the sea wires to produce twice as readable signals as rectangles within the limits of accuracy of audibility measurements, the last column showing the ratio of improvement over rectangle is obtained by multiplying by two the ratio of improvement over sea wires. Four cases are included in this table where unbalanced the signal was inaudible, giving a theoretical improvement in ratio of infinity, which, of course, simply means that it was not possible to obtain the audibility of the signal with the means at hand.

It is a little hard to say exactly what ratio of improvement is obtainable with the balanced system, the same depending so much on the character of the strays, but the writer believes that a ratio of improvement of 4.3-to-1 over the sea wires, that is to say, 8.6-to-1 over the rectangle, is a conservative estimate. As long as only one balanced set was in operation, much larger ratios of improvement were obtained. In fact the ratio of improvement averaged nearly twice as much. There was, however, an inevitable reaction when six sea wires were laid side by side within two hundred feet (61 m.), no one of the balanced circuits showing the degree of improvement that it showed when used alone. This is a very important point, and if new stations of this character were to be designed, it would be highly advisable, as recommended in one of the Belmar reports to the Bureau of Steam Engineering, to space the wires as far apart as possible. It was estimated in this report that two hundred feet (61 m.) between the different collecting systems would be a satisfactory spacing. At Belmar, the rectangles used in balanced work were entirely too close together, four rectangles being erected within a distance of three hundred feet (92 m.). When the Belmar system is compared with other systems on the basis of the work done in the summer of 1918, the important fact should be borne in mind that from four to six balanced circuits were in continuous operation in close proximity to each other during the entire summer. Further evidence of the improvement produced by the advent of balanced sets at Belmar in the early part of April is shown in curves numbers 1 to 5 inclusive. Curve 1 represents the signals from Carnarvon at 14,000 meters. The vertical

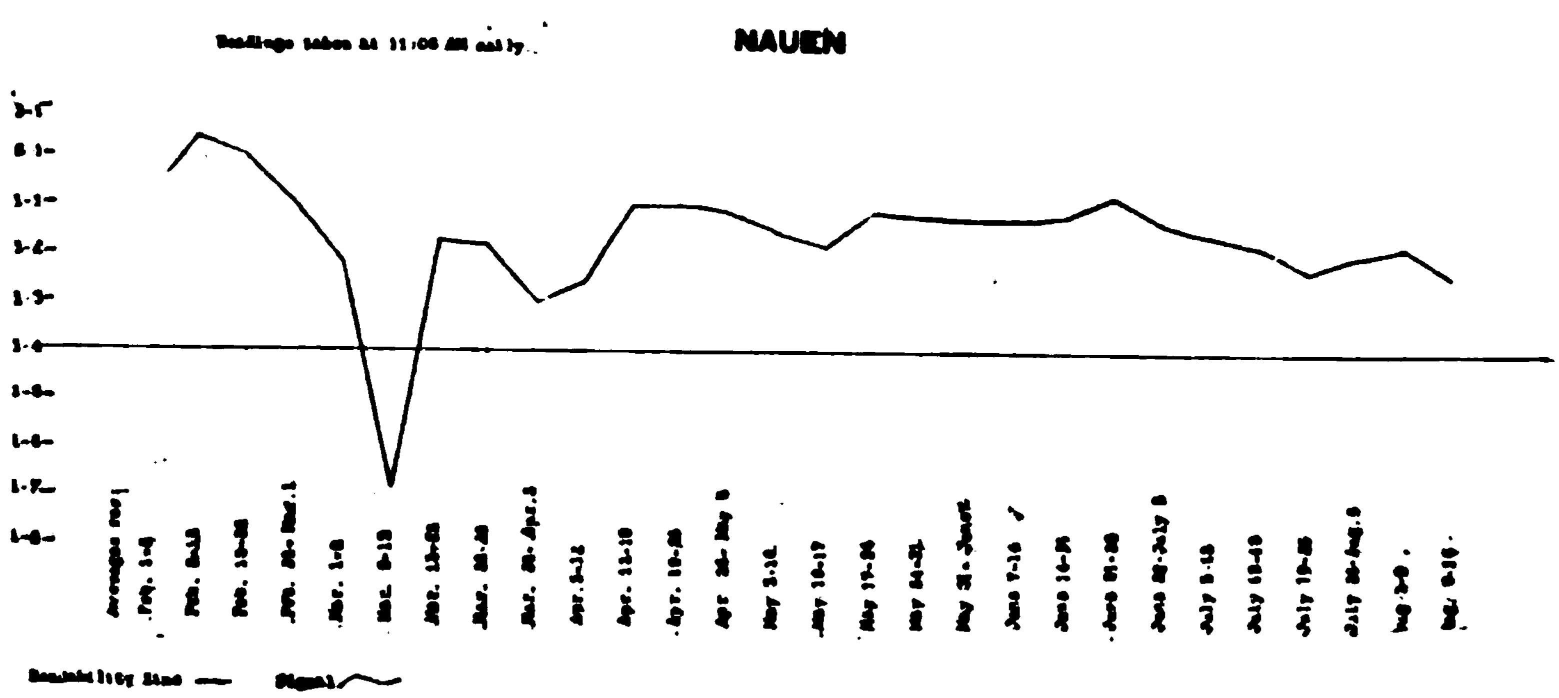
Date July	Station	Time P. M.	Balanced		Unbalanced		Ratio of Improvement over	
			Strays	Signals	Strays	Signals	Sea Wires	Rectangle
22	Nauen	3:30	200	80	400	40	4	8
22	Carnarvon	3:30	300	100	600	20	10	20
23	Nauen	3:30	160	20	600	Inaudible	Infinity	
23	Carnarvon	3:40	Signals inaudible on balanced and unbalanced sets					
24	Nauen	3:30	120	20	—	Inaudible	Infinity	
24	Carnarvon	3:40	250	40	—	Inaudible	Infinity	
25	Carnarvon	3:30	200	25	—	Inaudible	Infinity	
25	Nauen	4:00	80	20	80	5	4	8
26	Carnarvon	3:30	120	20	300	10	5	10
26	Nauen	3:40	80	25	800	80	3	6
27	Nauen	3:30	160	40	200	30	2	4
27	Carnarvon	3:40	100	20	300	30	2	4

ordinates are readabilities, that is to say ratio of signals to strays, the base line representing an adverse ratio in favor of strays of 4-to-1, which the writer believes is the practical limit of readability. It will be seen that from February 22nd to



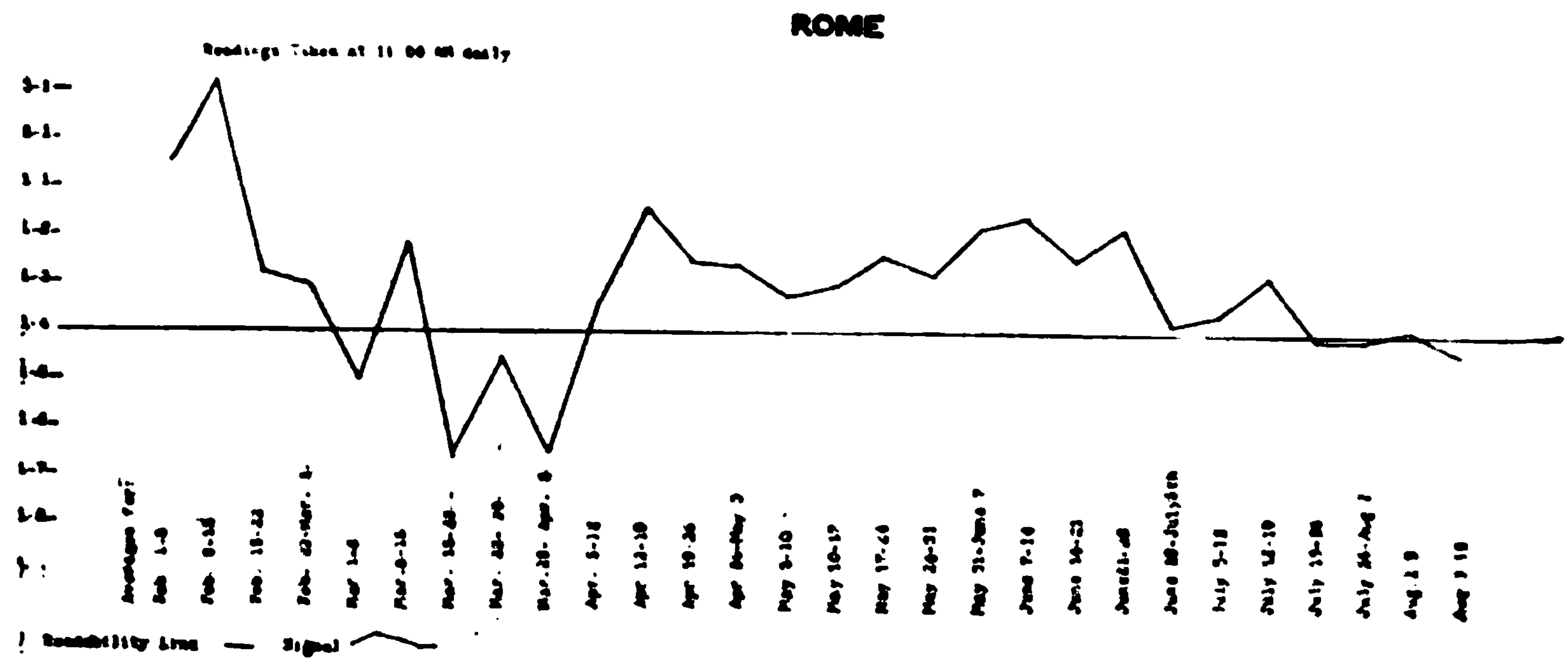
Curve 1

March 29th, Carnarvon's signals were getting very bad, being unreadable a large part of the time, but with the advent of the balanced system, they quickly jumped above readability and remained so during the bad summer months. After May 7th, Carnarvon shifted his schedule to the afternoon, which also produced an improvement in his readability, nevertheless without the balanced system he was usually unreadable during the summer months. Curve 2 shows a similar graph for Nauen,



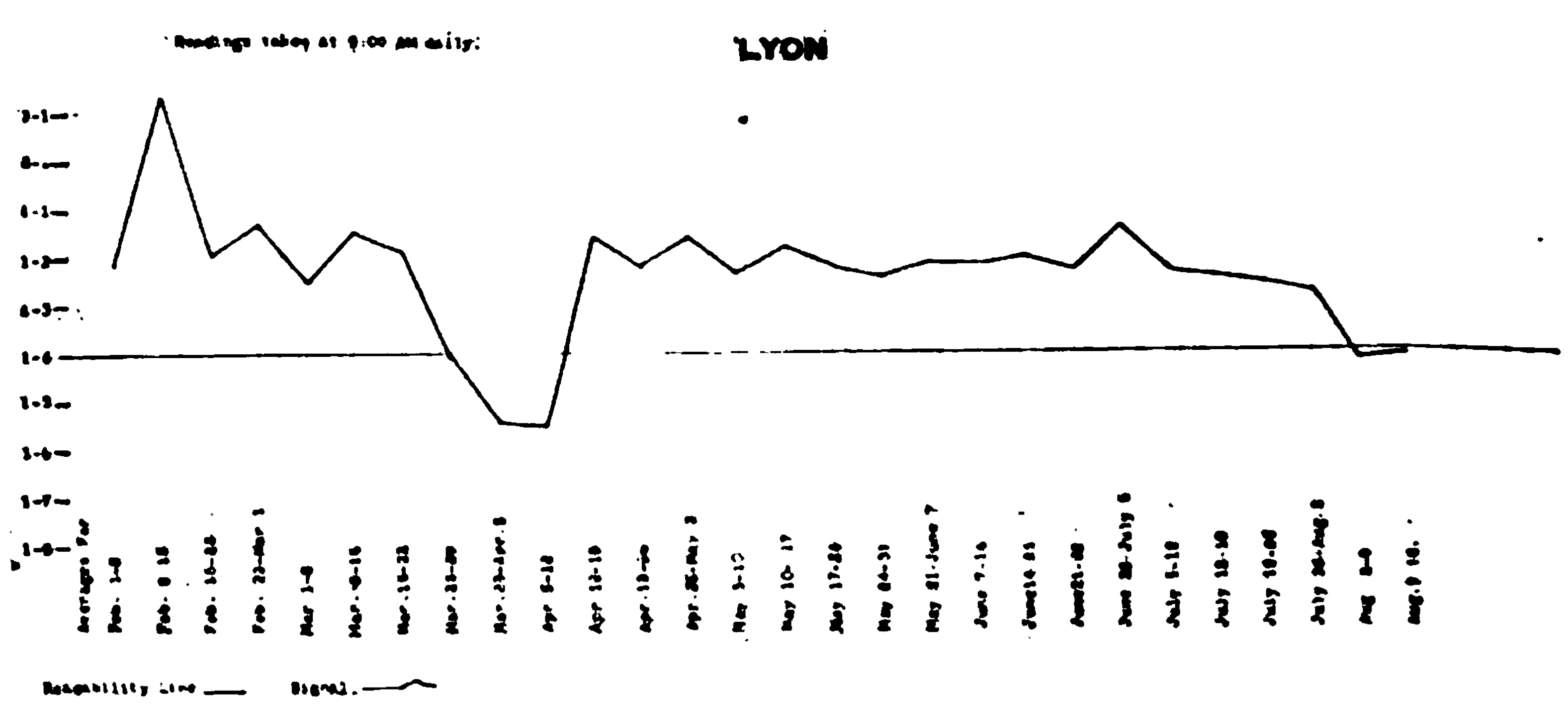
Curve 2

observations being taken on 12,600 meter wave. Curve 3 shows the improvement in Rome's signals on 11,000 meters after the first week in April, when the circuit was provided with a balanced system. The Lyons circuit was not provided with a



CURVE 3

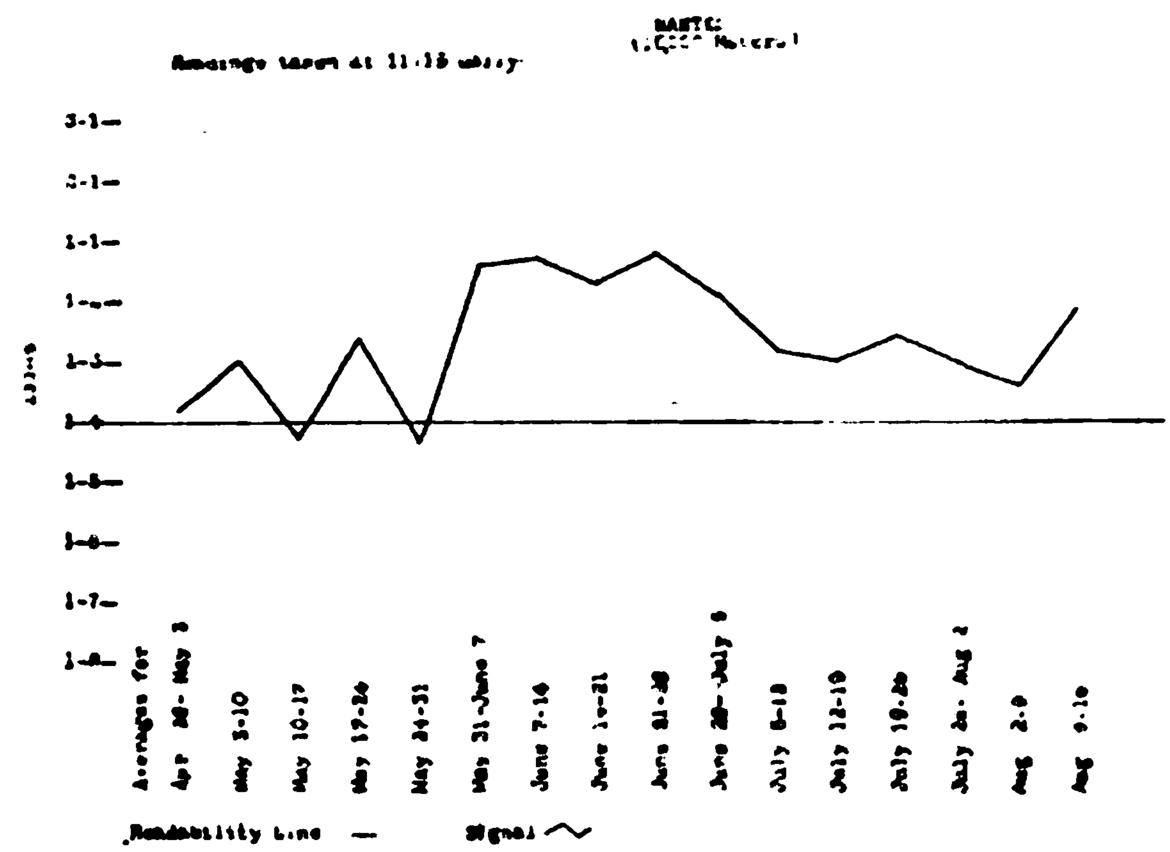
balanced system on account of the lack of a large enough rectangle for 15,000 meters until the second week in April. Curve 4 shows the resulting improvement. Curve 5 shows data ob-



CURVE 4

tained on signals from Nantes on 10,000 meters, but does not antedate the balanced system as we had no schedule with Nantes during that period. It is felt that the results at Belmar would have been somewhat better if the bluff across the inlet had not shielded the sea wires considerably. That this was the case

was demonstrated by experiments made with an 800-foot (244 m.) wire which was towed about further out in the inlet, receiving being done on board a small motor dory. The signals were noticeably stronger further away from the bluff.



CURVE 5

8. DIRECTIVITY

It is difficult to carry out accurate experiments on directivity of ground wires. Perhaps the most conclusive experiments at Belmar were those referred to in the preceeding paragraph, where the 800-foot (244 m.) wire trailed from a motor boat was swung about in different directions, maximum signals always being obtained when the wire pointed either toward or directly away from the European stations. Work done by Mr. H. H. Lyon at New London indicated some difference in the strength of signal when the wire pointed toward the station as compared to what it was when the wire pointed away from it. This was not confirmed at Belmar. It is believed that Mr. Lyon's results may have been influenced by shielding effects, which was certainly not the case at Belmar, as the work was carried on well out in the inlet and away from the shore. Concerning directivity, it has been noted that the directivity does not seem as sharp on very distant stations as it does on nearby stations. This is, of course, fully understandable. There are many routes by which a signal might reach a station from a transmitter at a great distance. The most striking case noted at Belmar was where signals from Cavite, Philippine Islands, 8,300 miles (13,-

280 km.) distant, were received with an intensity loud enough to be heard with the telephones on the table, altho the shortest great circle line from Belmar to Cavite, which passes up thru Hudson Bay, the Arctic Ocean and thru part of Siberia, strikes Belmar almost exactly at right angles to the wires which were used for reception. In agreement with the lack of directivity at times on long distance stations stand certain results obtained by the writer at the Naval Aircraft Radio Laboratory with very long wave direction finders. There are times, especially around sunset, when the signal may appear to come from a direction many degrees different from the true bearing of the station. Signals have been obtained from New Brunswick, distant less than 200 miles (320 km.), which showed a variation around sunset of 68° in fourteen minutes, returning to normal bearing after sunset, rather large deviations persisting, however, thruout the evening. This matter will be reported in the "Bulletin of the Bureau of Standards," and is fully explainable on the basis of the refraction and reflection theory. Evidently on very long waves coming from great distances there will be times when the resultant wave front really comes from a very different direction than the geographical bearing of the station. It may be confidently asserted, however, that the directivity of the ground wires is of the same nature as that of closed loops, and that either one of them will show peculiar directive properties on very long waves. As long as the ground wire and the rectangle are arranged to have the same line of directivity, the balanced system works satisfactorily.

9. Balanced System Where Sea Wires Are Not Available

In this case the balanced system will not handle the strays well unless the ground wire is laid in moist ground so that it has a better stray ratio than the opposing collector. The opposing collector should be a rectangle, but might be a ground wire laid in dry soil, if such is available. The focussing property of the system might, however, be extremely valuable of itself, irrespective of the question of stray elimination. Results obtained at Tuckerton with the balanced system showed very favorable results on both points, altho the elimination of strays was not as good as that obtained at Belmar. When the set itself, which was placed 1,800 feet (549 m.) from the base of the Tuckerton tower, was shielded from direct influence of Tuckerton's radiation, remarkably excellent results in the way of eliminating interference from Tuckerton, were obtained. The ex-

periments were never completed, owing to the fact that the writer was detached as trans-Atlantic Communication Officer, while the Tuckerton experiments were only just begun. The essential feature of the balanced system seemed to be two collectors which have fundamentally different ratios of signals to strays. The writer believes that it may be possible to construct a collector which will imitate the properties of sea wires and permit the balanced system to be installed almost anywhere. The experiments with lead-covered cable, while by no means successful in this regard, showed, however, very definite progress in the right direction. The balanced system herein outlined has the disadvantage of requiring special properties for one of the collectors, which so far have only been obtainable with sea wires and with wires laid in moist soil. It has the advantage, compared with some other systems, of requiring a comparatively small extent of territory for its installation, the total distance from the rear of one of the receiving rectangles to the outer end of the farthest sea wire at Belmar being 1,800 feet (549 m.). Trans-Atlantic work can be very satisfactorily carried on with a total distance from one end of the system to the other of 1,400 feet (427 m.)

Summary

1. Ground wires show positive indication of optimum length only in the case of the lead-covered cable lying on the surface. It is highly probable that an optimum length also exists where the wires are laid in dry soil. There is a decided possibility of optimum length existing for wires in fresh water, but wires in salt water show no indications whatever of optimum length.

2. The optimum length, in any event, is much less sharply marked for long waves than for short ones.

3. Multiplex reception is possible on the same ground wire, water wire, or surface cable, it being possible to receive any number of long wave stations on different wave lengths simultaneously as long as the local oscillations are so adjusted that the sets do not heterodyne against each other within the range of audibility.

4. Wires laid in fresh water or in dry ground show the same ratio of signal to stray as a large rectangle. Lead-covered cable shows a better ratio of signal to stray than the rectangle, but the difference is not very great. The best ratio of signal to stray is obtained with wires laid in salt water; the next best ratio with wires laid in wet earth.

5. Wires in the earth have been laid as deep as seven feet

(2.1 m.) in dry ground without showing any diminution in signal strength, in fact the signal strength was greater than for wires buried two feet (62 cm.) deep. Wires may be laid in fresh water up to sixty feet (18.3 m.) and still receive excellent signals, but in salt water the signals fall off rapidly with the depth, satisfactory signals for trans-Atlantic work not being received below four feet (1.24 m.). From six to eighteen inches (15.2 to 45.7 cm.) is preferable for salt water.

6. Various experiments looking to the improvement of the receiver in the matter of eliminating strays have been described

7. A successful method of more or less completely eliminating strays before they reach the secondary of the receiver has been described in detail. Two types of balances have been utilized in trans-Atlantic work at Belmar. One depends upon the dissimilar properties of a land wire and a sea wire and the other upon the dissimilar properties of a sea wire and a rectangle.

8. The method of adjusting the Taylor balance system has been described.

9. The character of strays as handled by the Taylor system has been briefly discussed.

10. It has been pointed out that when a number of balanced circuits are operated in close proximity, the perfection of balance is somewhat impaired.

11. The ratio of improvement in readability is conservatively estimated to be 8.6-to-1 as compared to the rectangle, when six balanced circuits were used within a small radius at the same station.

12. Curves have been presented showing the marked improvement in receiving conditions at Belmar upon the advent of the balanced circuits.

13. The directivity of the balanced system has been discussed and is believed to be essentially the same as that of the receiving loop.

14. The advantage of the Taylor system in requiring a small area for its installation has been indicated and its possibilities outlined for stations where sea wires are not available.

15. The focussing effect of the system has been described and its value in connection with distant control work and the elimination of interference has been pointed out.

SUMMARY: The question of optimum length of (buried) ground wires for reception at a given wave length is discussed, and experimental data are given. The signal strength obtainable on such wires under various conditions is considered.

The signal-to-stray ratio on ground wires as compared to that on loop

receivers (rectangles) is found to be more advantageous, particularly under carefully chosen conditions.

After considering a number of methods of reducing strays which have already penetrated into the receiver circuits, there are described a number of more effective methods for reducing strays before their entrance into the receiver. Thus strays can be balanced out by using a sea wire and a land wire as opposing collectors, with an adjustable-phase differential coupling of some sort. The wiring of the arrangements used and the practical adjustment are given, together with some of the experimental results obtained therewith.

The distance of origin of strays (or interfering signals) can be adjusted for in balancing these out, so that an interesting "focussing effect" is obtainable wherein the effect of a stray (or signal) in the receiver depends on the distance to its source.

The ratio of improvement in readability of signals thru strays obtained by the above arrangement is given conservatively as 8.6-to-1.

Certain remarkable variations in directional effect sometimes obtained are then discussed.

AN OSCILLATION SOURCE FOR RADIO RECEIVER INVESTIGATIONS*

By

Julius Weinberger and Carl Dreher

(Research Department, Marconi Wireless Telegraph Company of America)

1. Introduction

The purpose of this paper is to describe a source of sustained oscillations for use in connection with radio receiver investigations, which has been found practical in the Research Department laboratories. A source was required that would reproduce the effects occurring in the reception of electric waves on an antenna, and which would also produce a minimum amount of mutual interference where several investigations were being carried on in the same room, along similar lines. The oscillating audion has been extensively used as such a source, as described by Austin[1] and Armstrong,[2] but the methods of these authors have some disadvantages when applied to our particular case. Firstly, the usual audion source gives rise to rather widespread electric and magnetic stray fields, even when the bulb is of the smallest available type (a receiving tube) and a minimum amount of power output is used, so that serious interference results when several investigations are being carried on in the same room on the same range of wave lengths. Secondly, in order to measure the energy input to the receiver, crystal detectors and galvanometers were used to indicate the received current; these require laborious calibration against a thermo-couple, which must be repeated if the crystal should get out of adjustment, and furthermore, the lower limit of current which can be thus measured is of considerably greater magnitude than that obtainable in long distance (e. g. trans-oceanic) reception. In the source which we shall describe, both of these difficulties have been avoided; by

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* Received by the Editor, March 29, 1919.

[1] Austin, Proceedings of The Institute of Radio Engineers, volume 5, page 239, 1917.

[2] Armstrong, Proceedings of The Institute of Radio Engineers, volume 5, page 145, 1917.

suitable construction stray fields have been greatly reduced, and, instead of measuring the *current* in the dummy receiving antenna, an *emf*. of known amplitude is introduced. This is exactly what happens when the set is used under operating conditions; assuming that there is practically no reaction of the current flowing in the receiving antenna upon the passing electromagnetic field. This is probably the case with the usual heavily loaded, high resistance receiving antenna circuit, the induced emf. being proportional only to the field strength and to the height of the antenna. The use of a known emf. instead of a known current has the advantage that one may obtain as small a known emf. as desired, by mutual inductance methods; the known current method, however, is limited by the sensitiveness of the detector-galvanometer system (which permits of a minimum current of only about 6 microamperes being measured, according to some recent results of Dr. Austin).

The main source of trouble, however, which we had to eliminate, was the stray field produced by the usual oscillation source. A circuit assembled out of ordinary types of coils and condensers, connected to a receiving audion so as to produce oscillations, gave rise to so strong a stray field that it was impossible for two investigations on the same range of wave lengths to be carried on within twenty feet of each other, without mutual interference from the oscillation sources. A number of experiments were therefore first made to determine the origin of the various fields surrounding the source.

2. Preliminary Experiments

The circuit used for the production of oscillations is shown in Figure 1. Here, L is a single coil, with a double tap brought out from its mid-point and connected to the plate battery of the tube, C is a variable condenser of about 0.001 microfarads maximum capacity, and the vacuum tube is electrically equivalent to one of the usual small receiving types. Now, in this circuit each element—coil, condenser, and tube—gives rise to a stray electric field, and the coil also to a magnetic field. When oscillations are produced in this circuit, the point to which the filament is connected generally remains at a constant, or earth potential—probably because of the high capacity between the storage batteries which light the filament and earth—and the plate and grid terminals of the loop LC oscillate at high, radio frequency potentials; thus any metallic bodies connected to the latter points (such as the plates of the variable conden-

rise to stray electric fields, and these cause extremely powerful signals in a receiving set when the source is within several feet of it.

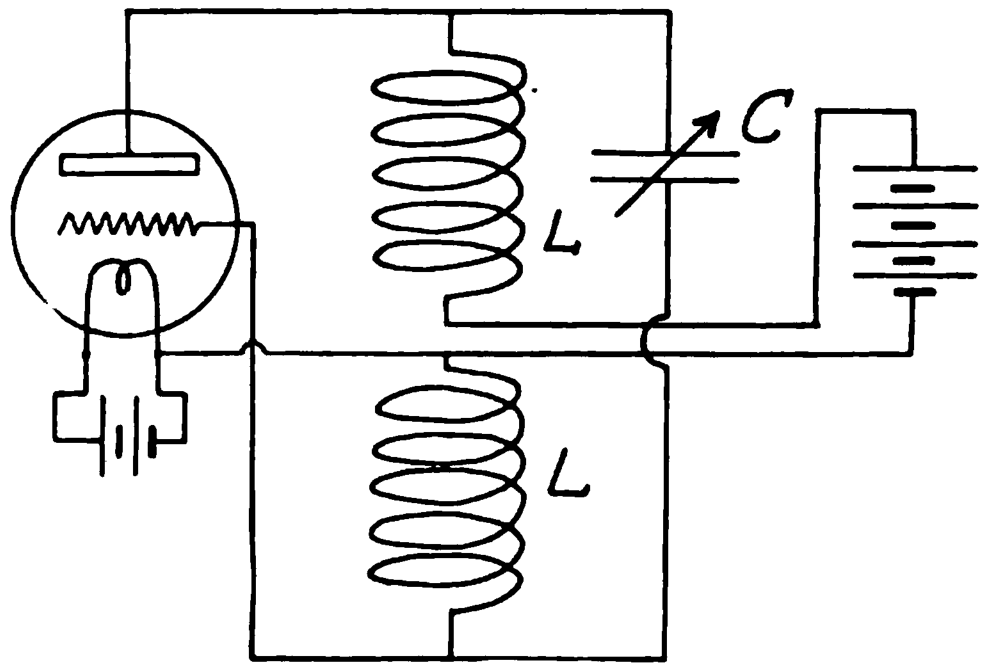

FIGURE 1

In order to study the exact manner in which the induction between the source and receiver took place, an elementary receiving set of the oscillating audion type was mounted on a board so that it could be moved about readily by the observer. A wiring diagram thereof is shown in Figure 2. The bulb was equivalent to the usual receiving type, L_1 and L_2 were single layer solenoids about 4 inches (10 cm.) in diameter having an inductance of 400 microhenrys each, C_1 a condenser of 0.0007 microfaiads maximum capacity, C_2 was 0.005 microfarads, C_3 was 0.001 microfarads and R about 500,000 ohms. B_2 was a battery of 20 volts and T a pair of Baldwin receivers. All the

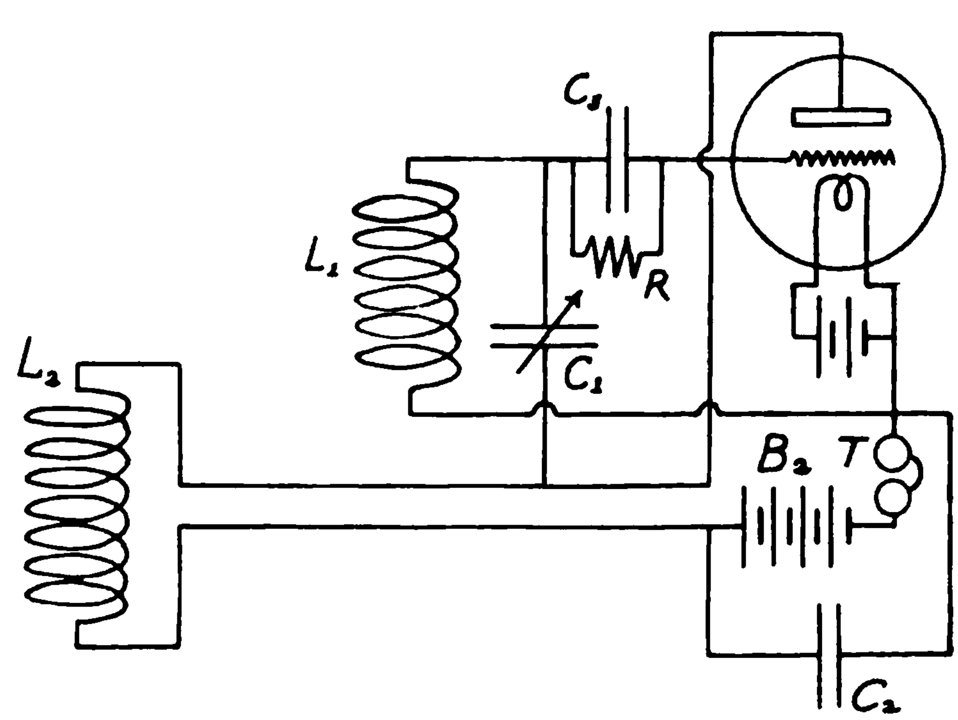

FIGURE 2

apparatus was fastened to a board except coil L_2, which was connected by a long pair of twisted leads to the rest of the receiving set so that it could be used as a search coil. The board containing the apparatus was usually placed at a sufficient distance from the source so that it was out of range of the effects of the fields therefrom, the search coil being used to pick up signals. When desired, the observer could handle the coil with a long wooden pole so as to eliminate effects due to electric fields picked up by his body.

The oscillator was set up as shown in Figure 1, the apparatus being the following: L, two single layer solenoids each about 4 inches (10 cm.) in diameter and each having an inductance of about 170 microhenrys, placed close together; C, a variable air condenser, of 0.0015 microfarads maximum capacity. A receiving tube was used, and the plate battery was 50 volts.

The search coil gave us a convenient means of distinguishing the kinds of fields around the oscillator, depending upon the following principle: If the coil is placed in a field which is purely magnetic, then a definite maximum and minimum signal is heard as the coil is revolved about its vertical axis, the maximum and minimum being 90 degrees apart. If the coil is placed in a field which is purely electric, then the signals remain of the same intensity all around, as the coil is revolved; inasmuch as the coil acts as a metal body upon which charges are induced. If both fields are present, then, as the coil is revolved, it is found that the maximum signal is not equally loud at two points 180 degrees apart. That is, if a maximum is obtained, and the coil then reversed, end for end, there is a distinct difference in loudness. This is because the emf. induced in the coil by the electric field has the same direction regardless of the reversal of the coil; but the emf. induced by the magnetic field is reversed relative to that induced by the electric field, when the coil is reversed. In other words, in one position of the coil the effects due to the two fields add, and in the other position they oppose. This phenomenon gives us a useful method of estimating the relative intensity of the two kinds of fields; we employed it to tell when we had only a magnetic field to deal with, or when both electric and magnetic fields were present.

(a) SHIELDING:

With the oscillator operating at a wave length of about 1,700 meters, it was found that just audible signals were obtainable within a radius of about 15 feet (4.6 meters); the signals

were due to an electric field entirely. The entire oscillation source (including filament and plate batteries) was next placed in a copper covered wooden box; the latter had its cover and one side on hinges, so as to permit access to the apparatus, and copper covering was also placed around the edges of these parts. If the box was closed completely, a fairly perfect copper shield was obtained around the oscillator, except for small openings where the junction between the shields on the hinged doors and on the sides of the box was not quite perfect (due to unavoidable kinks in the copper sheet). The importance of not having *any* such imperfections in the shield was not appreciated until later. The shield was always connected to one terminal of the filament battery.

With the box closed, just audible signals, due mainly to a *magnetic* field, were found about five feet (1.5 meters) from the shield. By moving the search coil about, it was found that the field was especially strong along the little openings between the doors and sides of the box, and that the field was predominantly electric. These openings were not very large—varying from a fraction of a mm. to 2 mm. Undoubtedly, lines of electric force issued from these cracks. Along the sides of the box, only a moderate magnetic field was found (due to the oscillator coils); it may be noted that the thickness of the copper sheet was about 12 mils (0.305 mm.), and at this wave length, the magnetic field passed thru very well. When the doors of the box were opened wide, signals were audible at the same distance as when the oscillator was out in the open, and were due to the electric field of the oscillator.

Another more perfect shield was constructed, consisting of a brass cylinder, with top and bottom plates of close-fitting heavy brass sheet. With this, it was found that the electric field was perfectly shielded off, only the magnetic field remaining, as before. Apparently the problem of shielding a radio frequency *electric* field is easily solved by the use of a perfectly closed screen.

In attempts to reduce the stray *magnetic* field, the coils of the oscillator were first surrounded by close-fitting iron boxes; the output of the oscillator, however, was found to be decreased so considerably by this measure, that no definite conclusions could be drawn as to whether the diminished external magnetic field was due to magnetic shielding or to the diminished output. Hence, the entire oscillator, in its brass case, was placed inside of a large galvanized iron can, and a galvanized iron cover placed

over it. The magnetic field came thru just as before. It may be concluded that galvanized iron does not act as a magnetic shield for weak radio frequency fields, either because skin effect causes eddy currents to flow in the galvanizing and the field is affected only by these eddy currents, or because of the low permeability of the iron for these fields. The latter is most probably the case, for, in connection with other work, we have noted that iron is a very poor shielding material for weak radio frequency magnetic or electric fields.

Another expedient for reducing the intensity of the magnetic field was the use of oscillator coils of small size and of the square or maxwellian cross-section type; with these, the magnetic field falls off in intensity very rapidly with the distance from the coil. Toroidal coils were also thought of, since these are generally supposed to have no external magnetic field.

(b) Multilayer Coils:

Two compact coils were built, each wound on a wooden form of the size shown in Figure 3. Each coil consisted of 80 turns of $7 \times 7 \times$ number 38 litzendraht (diameter of number 38 wire = 0.101 mm.), wound in five layers of 16 turns each. The induc-

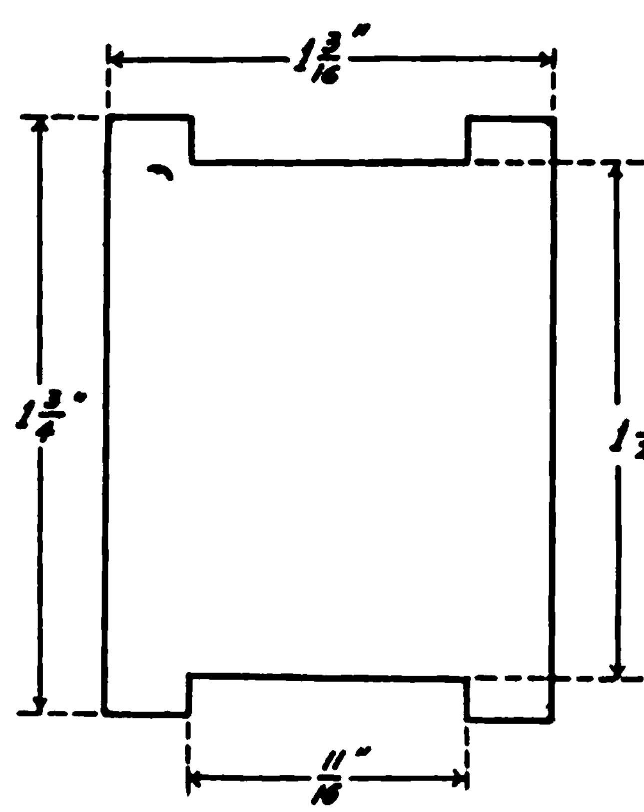

Figure 3

tance of each coil was 260 microhenrys. These coils were substituted for the single layer coils in the oscillator circuit, and the oscillating current was found to be about the same as before.

With the oscillator enclosed in its brass shield, just audible

signals were obtained at a distance of less than a foot (30 cm.) from the shield; the effect, as determined by the search coil was due only to the magnetic field of the oscillator coils, the electric field being as usual completely shielded.

(c) TOROIDAL COIL

A toroidal coil was wound on a wooden form of the dimensions indicated in Figure 4. It had 187 turns of 7×7×number 38 litzendraht, with a double tap at the mid-point. From the first turn to the mid-point constituted one coil, and from the mid-point to the last turn constituted a second coil. The inductance of both coils in series was 200 microhenrys. This coil was used in place of the multi-layer coils, and the radio frequency oscillating current found to be approximately the same as before.

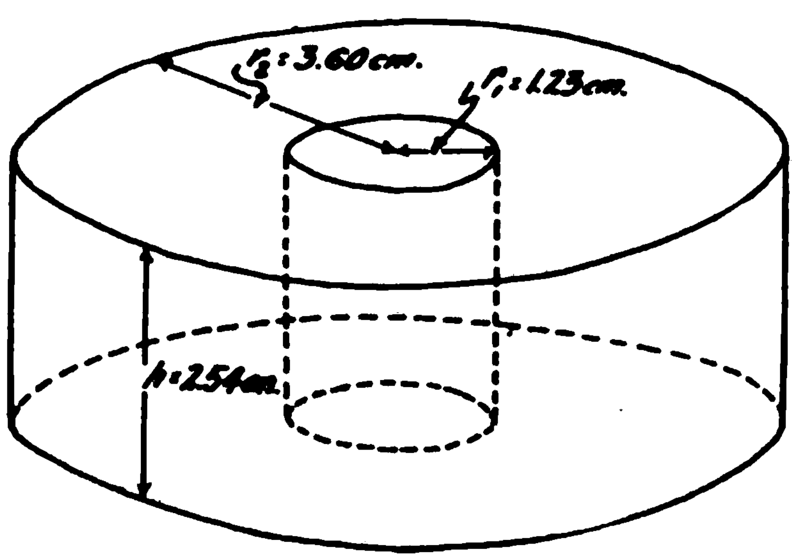

FIGURE 4

When the brass shield was closed, and the field investigated, much stronger signals were obtained than with the multi-layer coils. Rotating the search coil did not reveal any directional effects, the signals remaining of the same intensity as the coil was rotated. Such an effect could only be caused by an electric field.

However, we had already found that the shield screened off the electric field perfectly. We examined the field again, with the multi-layer and single layer-coils substituted for the toroidal coil and found no traces of an electric field—only of a magnetic field. Evidently the shield was still effective in cutting down the electric field.

Zenneck[3] describes a certain characteristic of toroidal coils

[3] J. Zenneck, "Elektromagnetische Schwingungen und Drahtlose Telegraphie," pages 53, 54.

which probably accounts for the peculiar results obtained above. As long as the current in the coil is constant, a magnetic field is present only within the coil; that is, for direct currents the toroidal coil has no external magnetic field. As soon as the current varies, this internal magnetic field also varies, and according to Maxwell's theory, an electric field is induced, with lines of force orthogonal to the original magnetic lines of force thru the coil, and terminating on the shield. The displacement currents flowing to the shield set up a second magnetic field, which gets thru the shield. The search coil, however, cannot be affected by this magnetic field (which is of the same shape as that within the toroid), because the search coil would have to be linked with the toroid windings in order to have the magnetic lines induce voltages in it. This varying magnetic field in turn induces in the ether another electric field, and that part of the magnetic field which exists outside the shield of course gives rise to an electric field in this region. It is this secondary electric field which affects the search coil in the observed manner.

We conclude, therefore, that the most effective arrangement of the source can be obtained by the use of compact, multilayer coils of restricted magnetic field, with copper or brass shields surrounding those parts of the circuit which give rise to electric fields, the shield being connected to one terminal of the filament battery.

(d) Effects of Bringing Leads out Thru the Shield

It will nearly always be necessary to bring leads out from points in the oscillating circuit, thru the shield; for example, it is inconvenient to keep the filament and plate batteries inside the shield, and it will also be necessary to bring out leads for coupling the oscillator to the receiver.

The filament and plate batteries were therefore located outside of the shield, connected by twisted leads, and the field was not found to be altered appreciably. It will be noted that the negative terminal of the plate battery is connected directly to the filament. If the plate battery were connected at any other point in the circuit, it would give rise to a strong stray electric field in case it were located outside the shield.

Next, a pair of twisted leads was brought out from point B, as shown in figure 5. Signals were immediately audible at a distance of 5 feet (1.5 meters) from these leads. Evidently a strong electric field was produced. But when the leads were brought out at point A, instead of B, the search coil had to be

brought very close to them in order to hear anything, and a purely magnetic field was found. This residual magnetic field produced by a twisted pair of wires was undoubtedly due to incomplete opposition of the magnetic fields of the individual wires. In this connection it may be noted that when a small

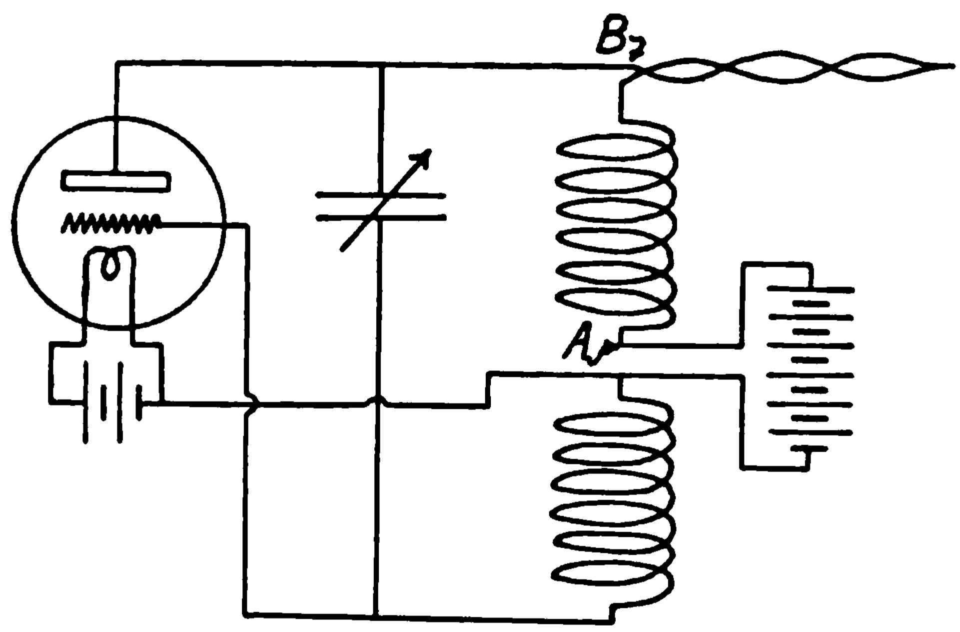

FIGURE 5

loop of perhaps one cm. in diameter was made by opening out the twisted pair, the increase in strength of the magnetic field of the wires was quite marked.

In may be concluded, therefore, that thoroly twisted leads may be brought out from a point in the oscillating circuit which is at constant (or ground) potential without materially increasing the existing external field. Such a point may be that immediately next to the filament, or, if the plate battery has its negative terminal connected to the filament battery, from the positive end of the plate battery. No leads should be brought out of the shield from other points in the circuit, or disturbing electric fields will be produced.

3. Description of Oscillation Source

Bearing in mind the conclusions from our preliminary experiments, the source described below was built. It is intended to give signals ranging from unit audibility to several thousand times audibility in a normal vacuum tube receiving set, over a wave length range from about 6,000 to 14,000 meters. It will be noticed that electric fields from the variable condenser, bulb and oscillating circuit inductance are shielded off; and that the energy from the oscillator enters the receiving set at only one point in the circuit of the latter, being controllable in amplitude

thru purely magnetic coupling. The influence of stray magnetic fields from the oscillator is negligibly small compared to that which comes in thru the path desired.

The wiring of the oscillator is shown in Figure 6, and photographs of it in Figures 7 and 8. Referring to Figure 6, the

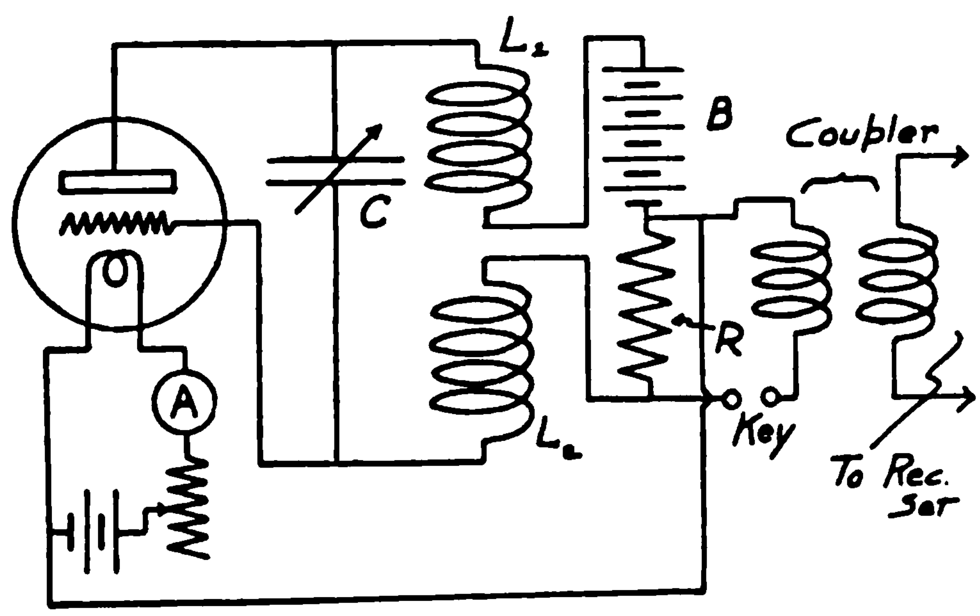

FIGURE 6

condenser C was of about 0.0007 $\mu f.$ capacity at its maximum, coils L_1 and L_2 were portions of a single coil having a double tap brought out at one point of the winding,[4] B was the plate battery which might be varied from 20 to 80 volts in order to vary the output, R was a resistance of 12,000 ohms (a Ward Leonard resistance unit was used), and the coupler was one having coils of about 40 microhenrys each, with a maximum mutual inductance of about the same value. The purpose of the particular arrangement adopted for sending signals necessitates explanation. It will be noticed that the resistance R, of 12,000 ohms, is inserted in the grid side of the oscillating circuit, and is shunted by the primary of the coupler in series with the binding posts marked "Key." (These may have a sending key or automatic transmitting arrangement connected to them, outside the oscillation source.) When the key is open, the resistance R effectually prevents oscillations from occurring; when closed, the resistance is short circuited by the primary of the coupler, and oscillations occur thru it. This particular system was found to be the only one which gave clear signals when reception was

[4] This coil was especially made to be as compact as possible. It consisted of 1,200 turns of number 28 S.S.C. wire (diameter of number 28 wire = 0.32 mm.) with the taps brought out at 300 turns, wound so as to have an inside diameter of 1.4 inches (3.56 cm.), an outside diameter of 2.6 inches (6.6 cm.) and a width of 0.6 inch (1.52 cm.). The total inductance was about 73 millihenrys, L_1 being 44 and L_2 5.3 millihenrys, respectively.

carried on by the heterodyne method; we had previously tried inserting the key in the plate battery leads and in other parts of the oscillation circuit, but the received signals always had a little variation in tone as the key was pressed or released. That is, as the oscillations were started and stopped, a slight variation in frequency to and from the normal oscillation frequency occurred, giving peculiar signals which were not clean-cut like the signals received from a radio frequency alternator or arc. It was found desirable not to interrupt the oscillation circuit completely, in sending, and the 12,000-ohm resistance cut into or out of circuit gave best results. In this connection, it may also be mentioned that good contact is very important in sending, with these small oscillating currents; the ordinary sending key when firmly operated in the manner customary with good operators, gave satisfactory signals. However, when an omnigraph was substituted for the key, so as to provide automatic transmission, the contacts were found to be insufficiently firm; hence, a telegraph relay (Western Union Company's type 3C signal relay) was used instead, which was itself operated by the omnigraph. The large contact points and firm pressure on the relay armature gave clean-cut signals; it was found desirable to connect the iron case of the relay magnet to the negative end of the filament to eliminate electrostatic induction from the interrupted direct current traversing the winding of the relay. Even this is not wholly satisfactory, however, and the best method of getting rid of clicks in the receiver from the relay direct current would undoubtedly call for thoro iron and copper shields completely around the magnets. We prefer to carry on tests with one man sending unknown material on a hand key, which has the double advantage of absence of direct current clicks and the reception by the other man of material unknown to him (one very soon becomes familiar with the material on some automatic transmitters).

Figure 7 shows the assembly of parts; individual units of commercial apparatus were merely assembled on a board. At the left, rear, is a filament ammeter, which is extremely necessary since the power output and frequency calibrations of the oscillator are correct only at one particular filament current; on the back of the ammeter is fastened a filament rheostat. At the right of the ammeter is the resistance unit R (shown better in the next figure); in front of it the variable condenser. The box-like affair in the middle is a wooden case covered with copper sheet, serving as a shield over the bulb and oscillating

Figure 7

circuit inductance; in Figure 8 the latter elements are shown with the cover removed, resting on a copper plate over which the copper-covered box fits tightly (the copper on the box being brought around under its lower edges so as to make contact with the copper base plate). At the extreme right is the coupler between oscillator and receiving set. The oscillation circuit is connected to the *rotating* coil of the coupler, the receiving set

Figure 8

to the *fixed* coil. This is found necessary, since the coil which connects to the receiver is subject to some magnetic induction from the oscillation circuit inductance unless it is suitably placed with respect to the latter; by using the fixed coil of the coupler for the receiver connection and placing this so that its windings are at right angles to the magnetic field lines of the oscillation circuit inductance (which lies horizontally), this difficulty is avoided. The exact position is easily found by trial; the leads from the oscillating circuit to the coupler are short-circuited, the receiver connected by twisted pair leads 3 or 4 feet (1 meter) long to the stationary coupler coil and the entire coupler rotated until no signals are heard. Then there is no magnetic induction between the oscillating circuit and receiver, and any magnetic induction thereafter will be due to the field from the moving coil of the coupler (which is in the oscillating circuit). Usually we have found that the proper location of the coupler is as shown in the photographs.

The variable condenser we used had a metal case (brass or aluminum), but had its plates suspended from an insulating top; we found that it was necessary to complete the shield by putting a copper plate underneath this top, with a small hole cut where the shaft of the moving plates came thru. The metal scale was *insulated* from the moving plates, and we found that a considerable electric field leaked out thru the hole where the shaft came thru unless this scale was grounded to the rest of the condenser shield (it will be noted from the wiring diagram of the set that neither set of plates may be grounded); hence a small phosphor bronze brush was arranged to bear against the moving scale and connect it to the shield. With these precautions no appreciable electric field was found to emanate from the condenser. The two leads to the condenser were brought out thru the shield over the bulb and coil, and were very short; it was not found necessary to shield them, altho we had expected that it might be required. All shields were connected to the negative end of the filament.

4. Results Obtained with Oscillation Source

(a) Calibration Data

In Figures 9, 10, 11, and 12, there are given certain data relating to the constants of the oscillator. Figure 9 gives the mutual inductance between coils of the coupler for various settings. This was made on an inductance bridge at audio frequency.

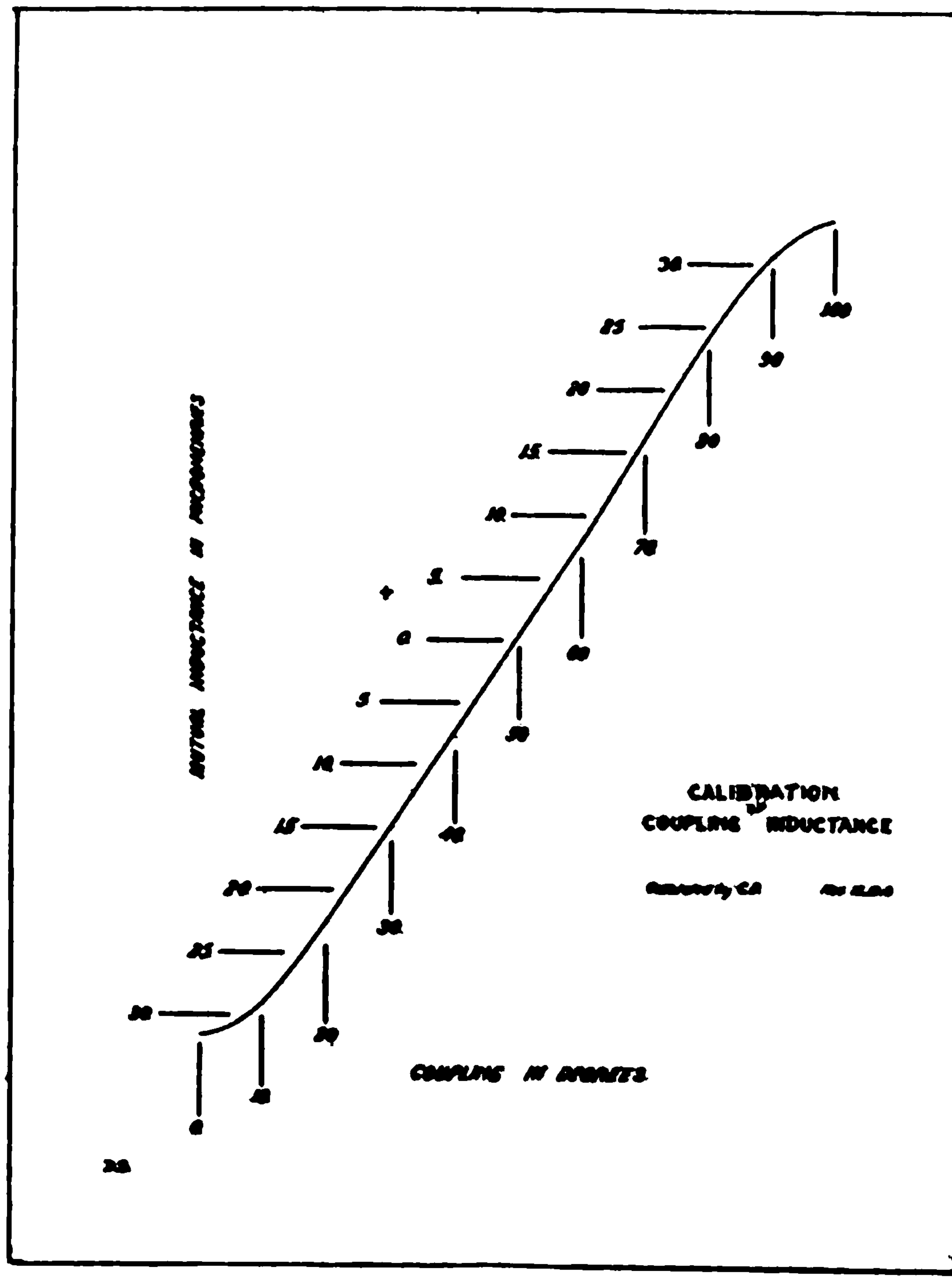

FIGURE 9

Figure 10 is the wave length calibration, which is correct only for the type of tube employed, operated at a filament current of 1.05 amperes; for other tubes and other filament currents this calibration (as well as the other calibrations given below) will vary slightly. However, when a given tube is always operated at the same filament current and plate battery, a calibration once made will remain the same for a long time. We have operated a tube day after day, at the same adjustments, and found absolutely no variation in the pitch of the *heterodyne signal* on a receiving set kept at constant adjustment. This calibration was made by receiving signals simultaneously from a large oscillator which had previously been calibrated by a wave meter and from our little oscillation source, on a crystal detector set. The interference between received currents due to the two sources gave rise to a beat tone in the receiver, and when zero beats were obtained the small oscillator was operating at the same wave length as the large one.

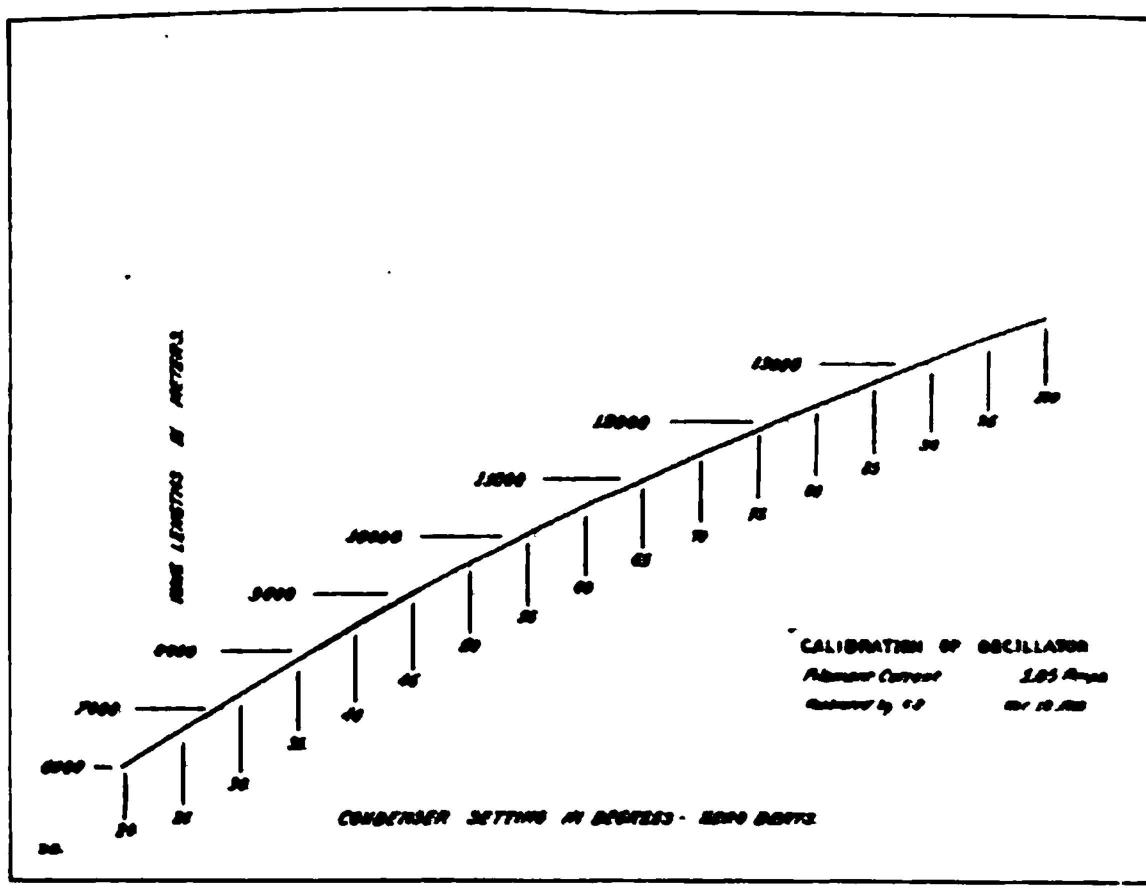

FIGURE 10

Figure 11 shows the radio frequency current in the oscillator circuit (that is, thru the primary of the coupler) with 20, 40, 60, and 80 volts plate battery. This was obtained by inserting a thermo-couple and galvanometer arrangement in series with the coupler primary as shown in Figure 13. The thermo-couple cannot be inserted directly in the circuit, since there is some

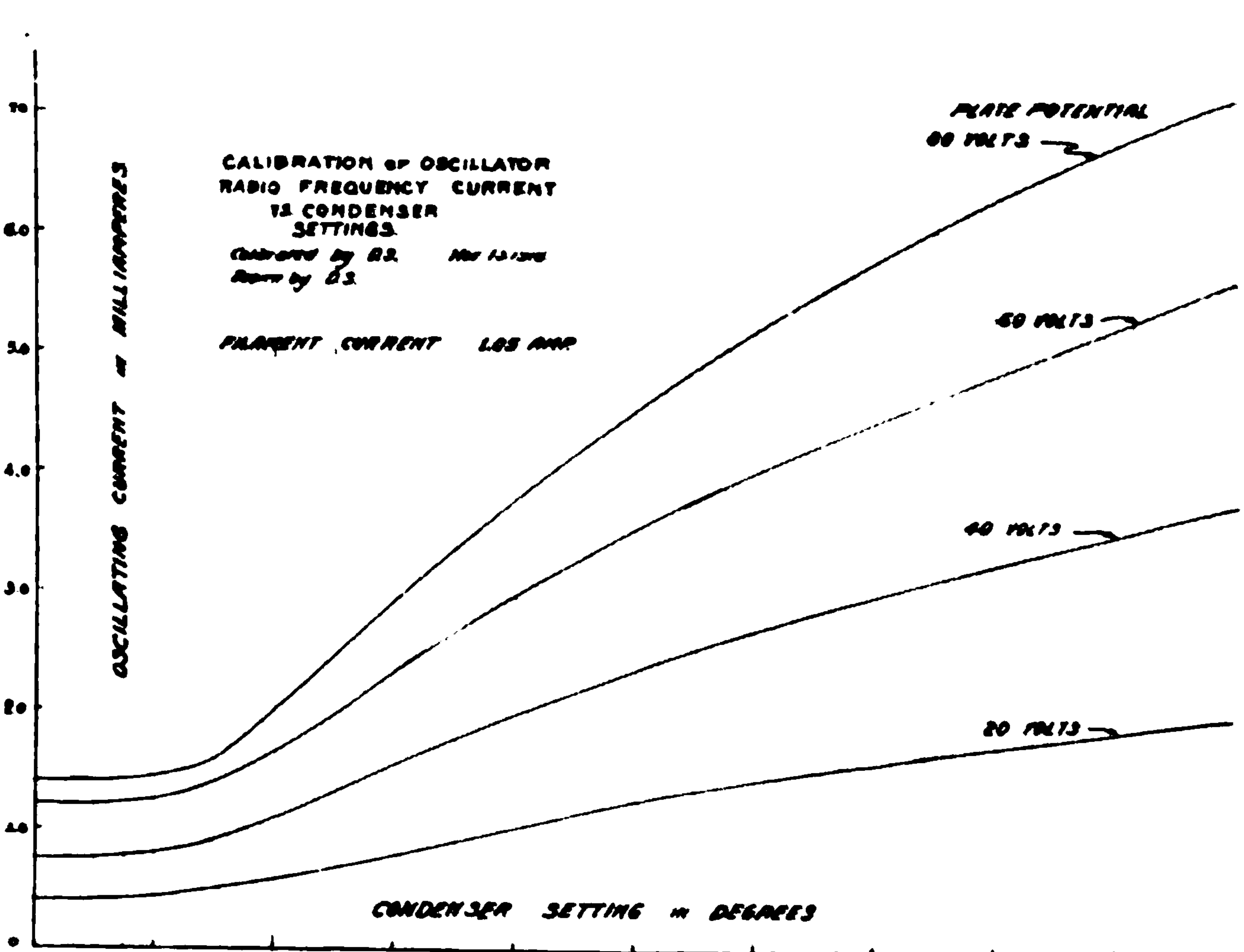

FIGURE 11

direct current flowing here; this direct current partially leaks thru the d. c. galvanometer and causes entirely erroneous deflections thereof. For example, reversing the circuit connections to the thermo-couple may cause different deflections, or even reversed deflections, to occur. Thermo-couples (even of the "heater" type), should never be connected directly in a circuit containing both direct and alternating currents; the arrangement shown in the figure is, however, free from this objection cited.

It will be seen that an inductance L, of impedance considerably higher than that of the thermo-couple heater, is shunted by a condenser in series with the heater. In our case L was 1 millihenry, C was 1 microfarad, the thermo-couple was one of R. W. Paul's vacuum types having a heater resistance of 0.8 ohms, and the galvanometer was a Leeds and Northrup suspension galvanometer (model 2285, 10 ohms coil resistance, 7 seconds period). The combination of thermo-couple and galvanometer alone had been previously calibrated at 60 cycles against electrodynamometer instruments.

Figure 12 shows the emf. induced in the stator of the coupler, at a wave length of 12,500 meters (this is the wave length of the station at Nauen, Germany) computed from the observed currents, mutual inductances, and frequencies.

$$e_2 = M \omega i_1$$

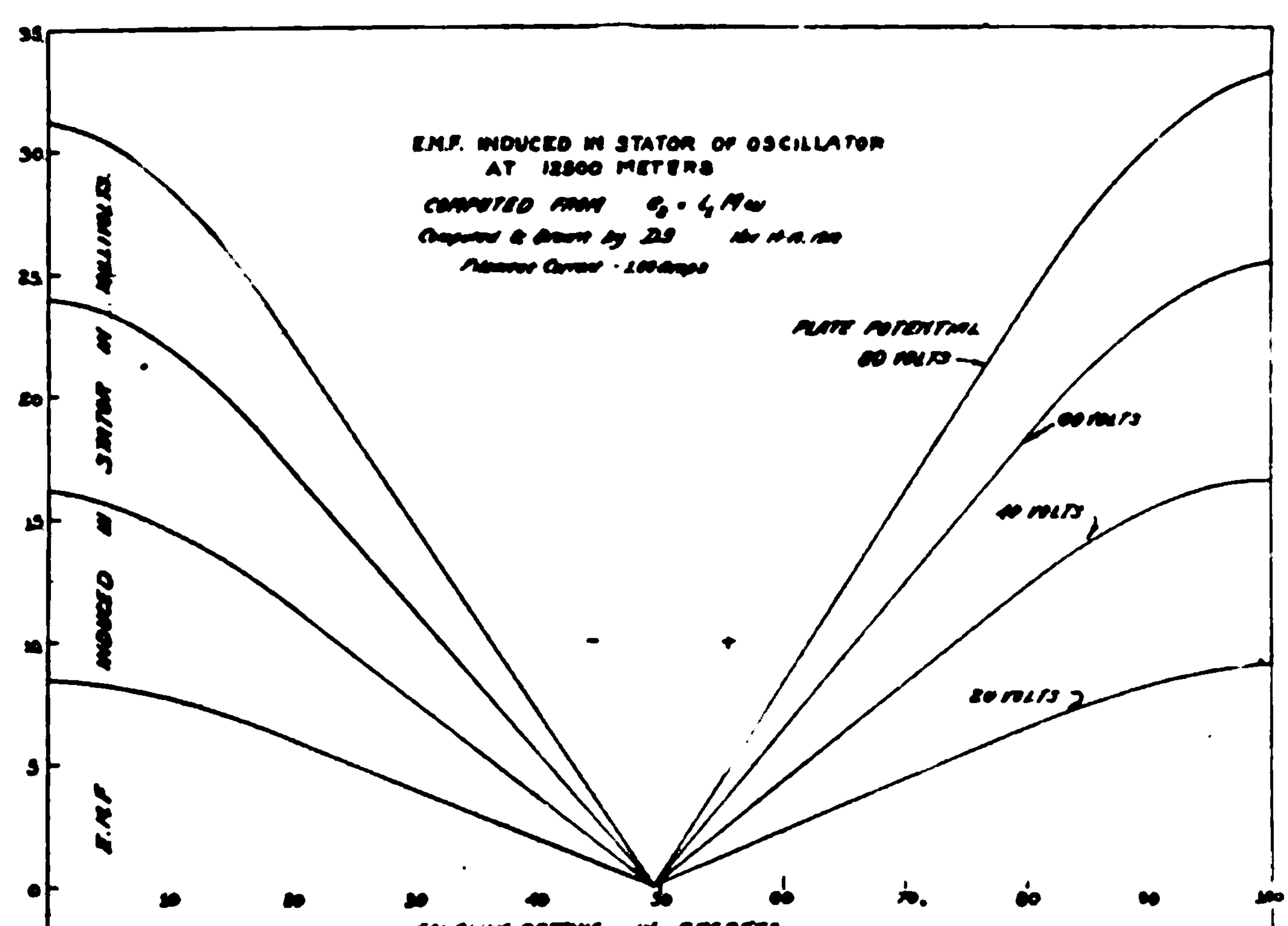

FIGURE 12

These values are given for various plate battery potentials. To find the voltages at wave lengths other than 12,500 meters, this equation may be used in connection with the curves previously given.

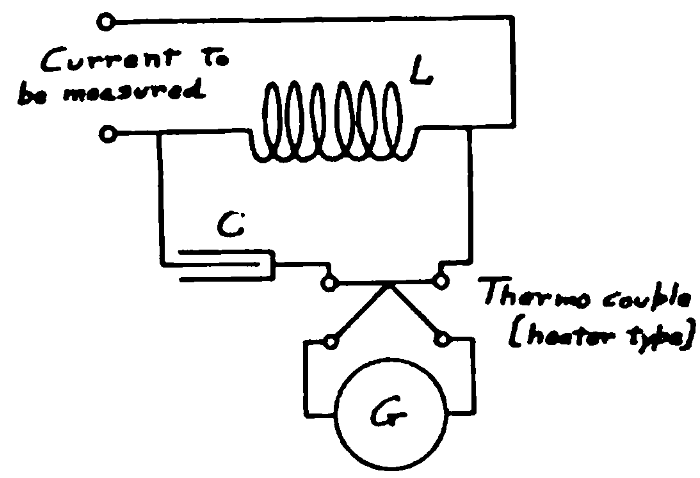

FIGURE 13

(b) AUDIBILITY OF SIGNALS COMPARED WITH MUTUAL INDUCTANCE BETWEEN COUPLER COILS

As a test of our assumption that the emf. induced in the stator of the coupler was proportional only to the mutual inductance between coils, the audibility of signals received by the heterodyne method on an experimental receiver was compared at various settings of the coupler. According to Dr. Austin's results, the audibility should be proportional to the emf. introduced into the receiver; and such proportionality was indeed found, the audibility being proportional to the mutual inductance, within limits of error in this class of measurement. In this connection, it may be observed that the customary method of measuring audibility in vacuum tube receivers, namely, putting two pairs of telephones in series and shunting one of them with the audibility meter, gives markedly incorrect results. If one observer listens in the unshunted pair of telephones, while the other is being shunted, the signals will be found to vary appreciably as the shunting resistance is varied. Obviously, the impedance of the output circuit of the tube is not maintained sufficiently constant by the extra pair of telephones. We found that to maintain the signals of the same intensity with or without a shunt on the telephone, it was necessary to add an inductance of about 10 to 15 henrys in addition to the customary extra pair of telephones, in the plate circuit of the tube. Such an inductance may be made of about 10,000 turns of fine copper wire on a silicon steel laminated core about

0.5 inch (1.25 cm.) in diameter and about 3 inches (7.62 cm.) long. A convenient method is to wind the wire on a hollow insulating tube, in which the core laminations may be slipped in and out and then, with the coil on an audio frequency inductance bridge, insert enough laminations to give the inductance desired. The current in the bridge should be as small as possible, so as to work the steel at a magnetizing force which falls in the region where the iron permeability is constant; the ordinary buzzer-driven inductance bridge is quite satisfactory in this respect.

It was also found that no signals were obtained at exactly the point of zero mutual inductance, on the coupler (49.5 scale divisions); this is an accurate check on the assumption that the induction between coupler coils was practically purely magnetic.

(c) Operation of Set

In using the apparatus, it is first placed at a distance of about 3 feet (90 cm.) from the receiver under test, the filament current and plate potential being set at the calibration values. The coupler is then set at the point of zero mutual inductance, the receiver connected by thoroly twisted pair to the coupler stator, and one man transmits signals. If any are heard in the receiver they are due to a residual stray magnetic field which emanates from the oscillating circuit coil and gets thru the shield over this coil (a certain amount of the magnetic field is cut out by the shield, due to eddy currents, but some comes thru). The receiver or oscillator is then rotated about until no signals are heard. When this has been accomplished, the emf. desired is produced by suitable settings on the oscillator.

One use for this equipment might be the following: If it were desired to measure the emf. induced in a receiving antenna by a transmitting station, day after day, very accurate results could be obtained by having the secondary of the coupler connected in series with the receiving antenna (while reception was going on), and having one man send, by hand, material similar to that being transmitted by the station; while the second man adjusted the coupler setting until the signals were readable equally.[5] It is obvious that the results obtained would be independent of personal equations, of errors arising due to daily variation in receiver adjustments and stray intensity, all of

[5] This idea is originally due to Messrs. R. A. Weagant and G. H. Clark, altho for a different purpose.

which enter into the present method of taking only audibility measurements.

We wish to express our indebtedness to Dr. Alfred N. Goldsmith for his valuable suggestions in connection with this work; and to Messrs. Sonkin and Ringel, of this Laboratory, for assistance with the experiments and calculations.

SUMMARY: **A source of long-wave, sustained oscillations, for use in connection with investigations on radio receivers, and providing standard controllable signals of the same character as those due to actual radio signals, is considered. Heretofore, such sources have had the disadvantages that they necessitated the measurement of small "received currents," and also, that considerable interference was caused by stay electric and magnetic fields emanating from the elements composing the oscillation circuit. The latter makes it difficult to carry on numerous researches in the same room on the same range of wave length; while the former cannot be accomplished with present-day apparatus for the minute currents occurring in trans-oceanic reception. The source described provides an emf. of known magnitude, rather than a current, which is exactly what occurs under operating conditions. In order to devise methods of reducing the intensity of stray fields from the source, a number of preliminary experiments are made and suitable methods of shielding and construction of oscillator elements are determined. A practical source utilizing these principles is described and construction and calibration data given. Finally, the use of this type of source in connection with transmission measurements on long distance radio communication is proposed, to replace the ordinary audibility measurements.**

ON THE DETECTING EFFICIENCY OF THE THERMIONIC DETECTOR*

By

H. J. van der Bijl, M.A., Ph.D.

(Western Electric Company, Incorporated, New York City)

1. Introduction

Altho the thermionic detector, also known as the audion detector, is now used extensively in the reception of radio signals, I have not come across any publication which describes a satisfactory method of determining its detecting efficiency in absolute units. There has been no satisfactory way of expressing what constitutes a good detector. In the early stages of development of a new device the lack of quantitative expression is perhaps not so severely felt. But the thermionic tube or audion has now passed beyond these stages. The work that has been done on the device in this laboratory has reached a stage where the tube is designed to have definite electrical constants depending on the purpose for which it is supposed to be used and which are determined from equations giving the relation between these constants and the structural parameters of the tube, such as the structure of the grid, its position relative to the anode and cathode and so on. Definite expressions for efficiency have thus become a necessity.

The structural equations were formulated by the writer, on the basis of an extensive series of investigations, with a sufficiently high degree of accuracy to meet practical requirements.

It was, therefore, a comparatively simple matter, when the Signal Corps required tubes for field and airplane radio work, to design tubes having the electrical characteristics that were necessitated by such uses, for example, low power consumption and satisfactory operation over wide ranges of filament and plate battery voltages. And all that was necessary was to strengthen mechanically the structure of the tube to withstand the contemplated rough handling in the field and the vibration to which

* Received by the Editor, April 9, 1919. Presented before The Institute of Radio Engineers, New York, May 7, 1919.

tubes are subjected on an airplane and to solve the problems of quantity production. It should be remarked that the tubes designed and manufactured before the war by the Western Electric Company and that have been used since 1914 as amplifying repeaters on long distance telephone lines, attained a higher degree of precision and uniformity of operation, than any tubes that were manufactured for war purposes. The reason for this is that on telephone lines the tube is usually inserted in the line at some intermediate place between the receiving and transmitting stations, the repeater stations being so arranged as to relay telephonic currents in both directions. It was therefore necessary for the tubes to have definite electrical constants to prevent their insertion in the line from causing an unbalance. Considering, furthermore, that on long lines several repeater stations are used and that sufficient distortion can be produced to make the transmitted speech unintelligible unless the tubes are properly designed and operated, it will become apparent that tubes to be used for relaying telephonic currents have to satisfy rather rigid test specifications. In radio telephony, on the other hand, the vacuum tube receiving sets, containing detector and amplifier tubes, are designed to transmit current only in one direction, and furthermore work directly into the telephone receiver. Requirements on such tubes need therefore not demand such close limits.

In the case in which the tube is used simply as a power amplifier, such as a telephone relay, methods of measuring the power amplification have been in use for many years. The most commonly used method consists in making a transmission test in which a current of about 800 cycles is amplified by the tube and attenuated by an artificial line of known attenuation constant. If the length of the line is adjusted to attenuate the current as much as it is amplified by the tube the degree of amplification can be computed from the constants of the line. This affords an extremely simple and rapid determination of the degree of amplification that a tube can give. (See Appendix.)

The detecting efficiency can, however, not be determined by such simple means, because here it is necessary to obtain a relation between the audio frequency power in the output of the detector and the radio frequency impressed on its input side. This determination is made difficult by the necessity of measuring the extremely small alternating currents involved in any attempt to carry out the experiments under conditions approaching those met with in practice. The currents to be measured

in the output range from 10^{-8} to 10^{-6} ampere. The use of hot wire instruments is, therefore, entirely out of the question. To overcome this difficulty the telephone receiver has been resorted to as a measuring instrument. This means has been made use of in the so-called "audibility method," which is, however, subject to serious limitations as will be shown below. The unreliability of the audibility method makes it quite unsuitable for standardization purposes. In seeking to minimize the psychological and physiological influences attending measurements made by the audibility method, I devised the following method which it is the purpose of this paper to describe. This method which allows of relatively easy measurements and a good degree of accuracy, has been in use in this laboratory for two years for the purpose of standardizing detector tubes, and as a laboratory method has given very satisfactory results. It can be, and has in this laboratory been applied in the study of detection with heterodyne or regenerative circuits as well as in the case of straight detection of modulated waves.

2. THE PRINCIPLE OF THE METHOD can be gathered from the schematic diagram shown in Figure 1. Modulated high (radio) frequency oscillations can be impressed on the detector at C. The input voltages ranged from a few hundredths to a few tenths of a volt, and can be measured with the Duddell thermogalvanometer G and non-inductive resistance r. In order to

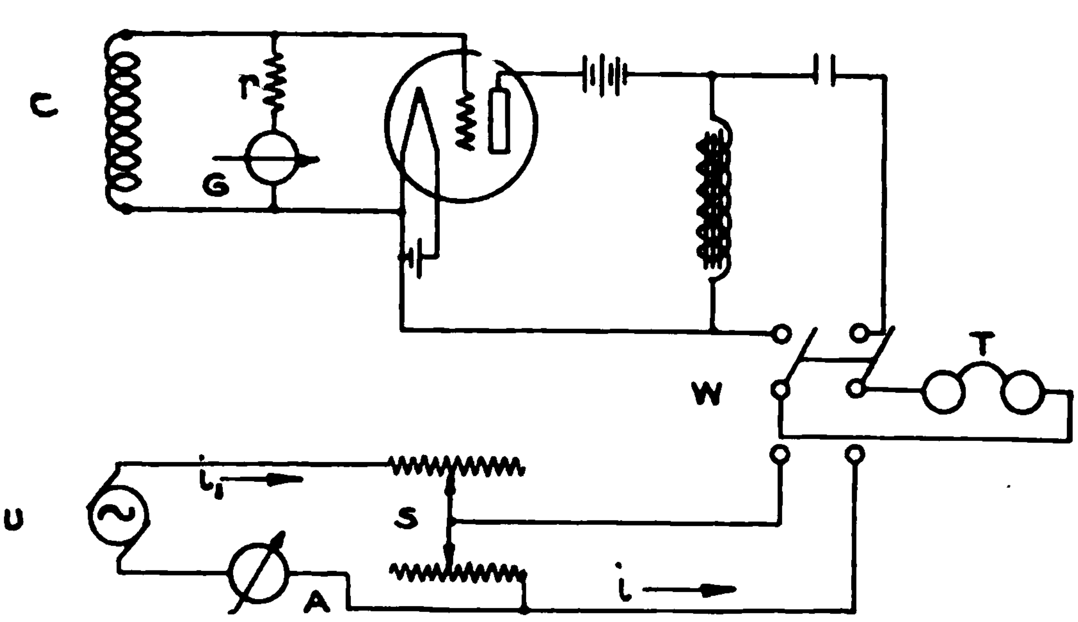

FIGURE 1

measure the resulting audio frequency current in the telephone receiver T, which we shall call the detecting current, a generator U is used to produce an auxiliary tone of the same frequency as that of the detecting current. The current, i_1, from the gen-

erator is so large that it can be easily measured with a thermocouple and milli-ammeter A, and can be attenuated by means of the shunt S, to a sufficient extent to make the branch current i equal to the detecting current. The current i can be computed in terms of the measured current i_1 and the constants of the shunt S, and this gives the detecting current when S is so adjusted as to make the note in the receiver of the same intensity for both positions of the switch W.

The shunt S is of the type commonly used in telephone measurements where it is known as a receiver shunt. It consists simply of shunt and series resistances so arranged in a box that when a telephone receiver of specified impedance is connected to one side the total impedance into which the generator works remains constant for all adjustments of the shunt. This nsures that the current i_1 remains constant for all values of i.

3. DETECTION COEFFICIENT. In connection with the study of thermionic detectors two of the main problems encountered are (1) the direct measurement of the detection coefficient, and (2) the formulation of the relationship between the detection coefficient and structural and operating parameters of the tube and circuit. The latter is of importance in properly designing the tubes while the former is important in that it furnishes a means of measuring and expressing the degree of merit of the tube used as a radio detector.

The relation between plate current and applied voltages can be expressed by

$$I = f\left(\frac{E_B}{\mu} + E_c + \epsilon\right), \tag{1}$$

where E_B and E_c are the plate and grid potentials with respect to the filament and ϵ a constant. The form of the function f depends upon the constants of the tube and circuit. It is important to discriminate between the characteristic of the tube itself and that of the tube and circuit. When the tube is used simply as a power amplifier its characteristic can be expressed as[1]

$$I = a\left(\frac{E_B}{\mu} + E_c + \epsilon\right)^2, \tag{2}$$

that is, in terms of the tube constants which again bear definite relations to the structural parameters. This equation is a first order approximation and consequently needs modification be-

[1] H. J. van der Bijl, "Phys. Rev.," 12, page 171, 1918; PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS, volume 7, page 97, 1919.

fore it can be applied to the tube when used as a radio detector, because detecting action depends on second order quantities, being determined by current rectification in the grid circuit when the tube is used with a blocking condenser in the grid circuit, and by the second derivative of the plate current characteristic when operated without the blocking condenser. For the present we shall consider only the latter case.

When we are concerned with the direct measurement of the detection coefficient, it is not necessary to formulate the characteristic in terms of the tube constants, since it is always possible to express the characteristic of the tube and circuit in a convergent power series:

$$J = a_1 e + a_2 e^2 + a_3 e^3 + \cdot \qquad (3)$$

where J is the varying current in the output, e an alternating emf. impressed on the grid circuit, and a_1, a_2 $\cdot \cdot \cdot$ are functions of the tube and circuit constants.[2]

The only term that need be considered as effective in detecting is the second, since the series converges so rapidly that all terms of higher order than the second can be neglected. We can, therefore, express the detecting current i as

$$i = a\, e^2 \qquad (4)$$

The quantity a represents what is referred to as the detection coefficient.

4. Relation Between Detection Coefficient and the Operating Plate and Grid Voltages. If the detecting current, i, be measured as a function of the "effective grid voltage" $\left(\frac{E_B}{\mu} + E_c + \epsilon\right)$, the input voltage e remaining constant, it will be found that as the effective voltage is increased (by increasing either E_B or E_c) the detecting current at first increases, reaches a maximum, and then decreases. This effect has doubtless been noticed by most workers using audion detectors.

If the parabolic characteristic equation (2) were correct to a second order the detecting current would not show a maximum but remain constant. The cause of this maximum lies in the potential drop in the filament occasioned by the filament heating current. It can be explained as follows:

The effect of the potential drop in filament on the characteristic of a simple thermionic value has been given by W. Wil-

[2] See J. R. Carson, Proceedings of The Institute of Radio Engineers, volume 7, page 187, 1919.

son of this laboratory[3] who finds that two characteristic equations are needed according as the potential difference V between the plate and (say) the negative end of the filament is less or greater than the potential drop E_f in the filament; thus

$$\left.\begin{aligned} I &= C\,V^{\frac{5}{2}} \text{ for } V<E_f. \\ I &= C[V^{\frac{5}{2}}-(V-E_f)^{\frac{5}{2}}] \text{ for } V>E_f. \end{aligned}\right\} \quad (5)$$

In order to obtain an indication as to where the maximum of detecting current occurs, without making an attempt to express the actual value of the maximum quantitatively, we can apply these equations to the three-electrode tube by substituting the effective voltage $\left(\frac{E_B}{\mu}+E_c+\epsilon\right)$ for V:

$$I=C\left(\frac{E_B}{\mu}+E_c+\epsilon\right)^{\frac{5}{2}} \quad (6)$$

$$I=C\left[\left(\frac{E_B}{\mu}+E_c+\epsilon\right)^{\frac{5}{2}}-\left(\frac{E_B}{\mu}+E_c+\epsilon-E_f\right)^{\frac{5}{2}}\right] \quad (7)$$

If the current is plotted as a function of the effective voltage according to equation (6) for all values of effective voltage less than the potential drop in the filament and according to (7) for effective voltages greater than the filament drop, the resulting values form a smooth continuous curve closely approximating a parabola. The second derivative, however, shows a distinct maximum at an effective voltage equal to the potential drop in the filament. The effect can be made clear if we consider the simple case in which the tube works in a non-reactive output circuit of resistance R. In this case the detection coefficient can be expressed as[4]

$$a=-\frac{1}{2}\frac{\mu^2 R_o R_o'}{(R+R_o)^3}$$

where R_o is the a. c. plate resistance of the tube and R_o' its first derivative. R_o and R_o' can be evaluated from the characteristic of the tube by the formation of the derivatives

$$R_o=\frac{1}{\frac{\partial I}{\partial E_B}} \text{ and } R_o'=\frac{\partial R_o}{\partial E_B}.$$

This gives, putting $\frac{E_B}{\mu}+E_c+\epsilon=V_e$:

[3] Paper read at the Philadelphia meeting of the American Physical Society, December, 1914.

[4] J. R. Carson, previous citation.

$$a = C'\left(\frac{R_o}{R+R_o}\right)^3 V_e^{\frac{1}{2}} \tag{8}$$

$$a = C'\left(\frac{R_o}{R+R_o}\right)^3 [V_e^{\frac{1}{2}} - (V_e - E_f)^{\frac{1}{2}}]. \tag{9}$$

according as V_e is less or greater than the potential drop E_f in the filament. It will be seen that for constant values of $\left(\frac{R_o}{R+R_o}\right)^3$, that is, when R is small compared with R_o, the coefficient a, which is a measure of the detecting current, increases with V_e according to equation (8) and decreases with V_e according to equation (9). The result is a curve like that shown in Figure 2. The simple rule, therefore, to obtain the best

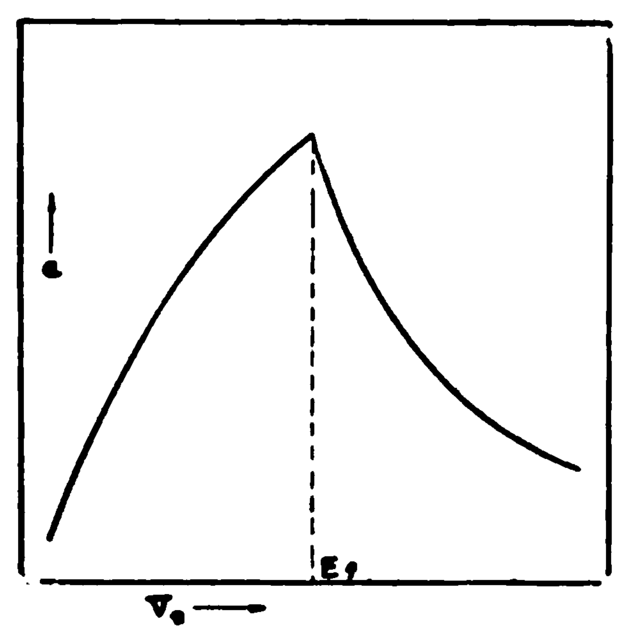

FIGURE 2

results when using a tube without a blocking condenser in the grid circuit, is to make

$$\frac{E_B}{\mu} + E_c + \epsilon = E_f. \tag{10}$$

Thus, supposing that $\mu = 12$, $\epsilon = -0.5$, and $E_f = 2.5$ volts, and the tube be operated without a grid battery, then the best plate voltage would be about 36 volts. An experimental curve is shown in Figure 3.

In determining the detection coefficient due regard must be taken of its dependence upon the applied d. c. plate and grid voltages.

5. INPUT SIGNAL WAVE. When measurements on the detection coefficient are made for the purpose of expressing the degree of merit of detectors it is necessary to specify the nature of the impressed radio frequency oscillations. The input volt-

age e may be characterized as the root-mean-square value of an unmodulated radio frequency voltage in which case a heterodyne local source of voltage or a regenerative circuit is needed to detect the oscillations; or it may be characterized as the root-mean-square of a modulated radio frequency voltage, which category would include spark signals.

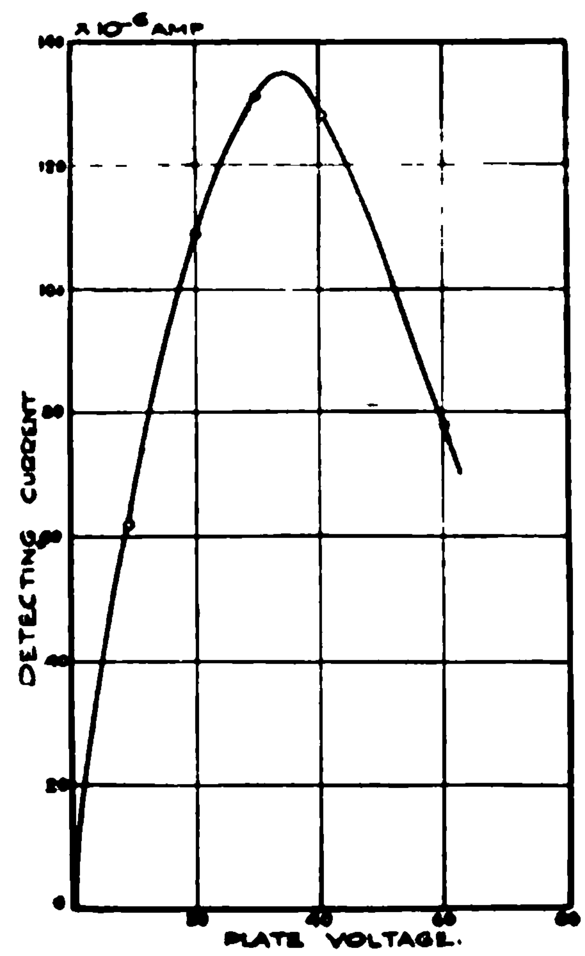

FIGURE 3

For the measurements under consideration, continuous unmodulated oscillations are unsuitable because the audio frequency response in the telephone receiver is determined mainly by the strength of the local auxiliary oscillations which is, in general, large compared with the strength of the received oscillations. Spark signals would be quite unsuitable, since here the character of the wave cannot be expressed in definite terms. The simplest and most easily reproduced type of wave is a radio frequency sinusoid modulated with an audio frequency sinusoid by means of a modulating device, such as the thermionic modulator, which enables the modulation to be controlled at will. This is the type of wave that was used in these measurements.

In order to specify fully the signal wave it is necessary to discuss briefly the process of modulation as effected by means of the thermionic vacuum tube. If we impress on the grid-filament circuit an emf.

$$e = e_1 \sin pt + e_2 \sin qt \tag{11}$$

where $\frac{p}{2\pi}$ and $\frac{q}{2\pi}$ are radio and audio frequencies respectively, then the current established in the plate circuit ("output circuit") can be represented by the convergent power series (3). In general all terms of higher power than the second can be neglected, so that the current in the output can be represented by:

$$J = a_1\,(e_1 \sin pt + e_2 \sin qt) + a_2\,(e_1 \sin pt + e_2 \sin pt)^2$$

Let the output of the modulator be tuned to a frequency range $p \pm q$, so that currents of these frequencies only will be radiated. We can, therefore, in evaluating the above expression, drop all terms representing frequencies that fall outside of the range $p \pm q$, such as $\frac{2p}{2\pi}$, $\frac{q}{2\pi}$, and $\frac{2q}{2\pi}$. A simple trigonometrical transformation then gives for the radiated wave:

$$A \sin pt\,(1 + B \sin qt). \tag{12}$$

where A and B are constants involving e_1 and e_2 and furthermore depend upon the constants of the modulator tube and circuit. This is a simple type of modulated wave and results from the curvature of the vacuum tube characteristic. Referring to Figure 4 which represents the plate current, grid voltage characteristic, if a radio frequency voltage $o\,a$ be superimposed on the negative d. c. grid voltage E_c, the output current will be proportional to $a\,b$. If E_c be reduced to E_c' or increased to E_c'', the output current will be increased to $a'b'$ or reduced to $a''b''$.

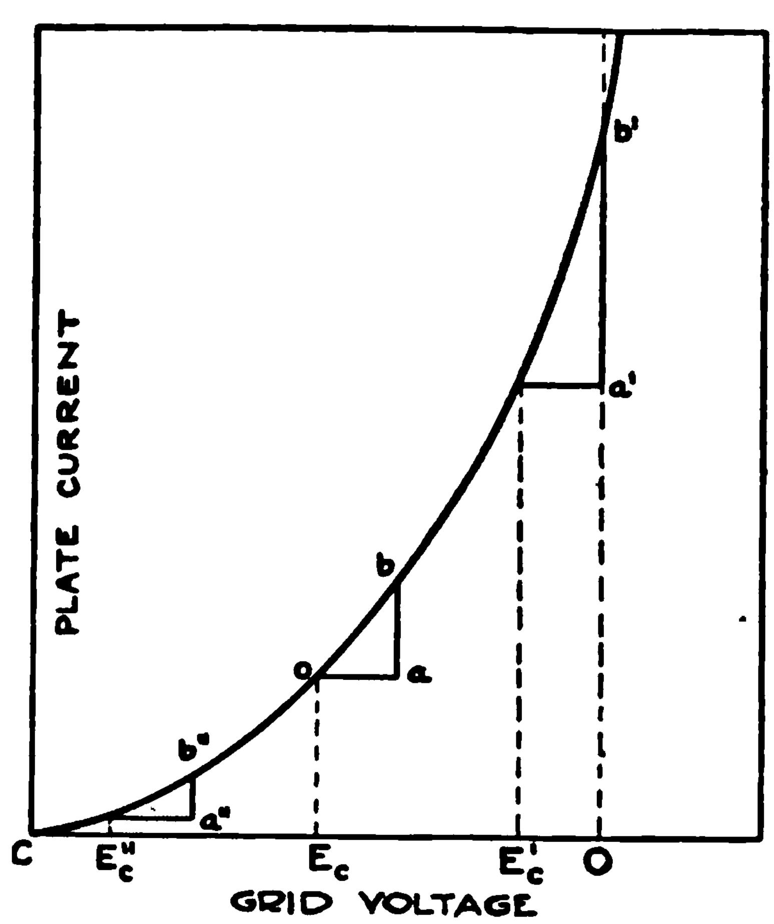

FIGURE 4

If this variation in E_c be effected by impressing a sinusoid of frequency $\frac{q}{2\pi}$ on the grid the output current will be given by equation (12) which represents a high frequency current the amplitude of which varies according to the low frequency $\frac{q}{2\pi}$. If the low frequency amplitude is equal to $E_c C$, the amplitude of the radio output wave will be reduced to zero whenever the low frequency input voltage attains its maximum negative value. In this case, B in equation (12) is unity and the wave can be said to be completely modulated.

It is very important in measurements relating to the detecting efficiency to insure that the input wave is completely modulated. The necessity for this can easily be seen from the following consideration.

If a voltage represented by (12) be impressed on the input of the detector, the detecting current can be obtained from:

$$i = a\,[A \sin pt\,(1 + B \sin qt)]^2$$

which on evaluating and dropping all terms representing inaudible frequencies gives for the detecting current which can be heard in the receiver:

$$a\,A^2 B \sin qt - \frac{a\,A^2 B^2}{4} \cos 2\,qt. \tag{13}$$

Its root-mean-square value is

$$X = a\sqrt{\frac{A^4 B^2}{2}\left(1 + \frac{B^2}{16}\right)} \tag{14}$$

The second term in the parenthesis represents a note of double the fundamental frequency and is generally weak enough to be neglected, so that we may write

$$X = \frac{a\,A^2 B}{\sqrt{2}} \tag{15}$$

The r.m.s. value Y of the modulated input represented by (12) can be evaluated by putting $p = n\,q$, since p is large compared with q. This gives

$$Y = \sqrt{\frac{A^2}{2}\left(1 + \frac{B^2}{2}\right)}. \tag{16}$$

Comparing (16) with (15) it is seen that for a constant value of the input Y, as measured by the Duddell galvanometer G (Figure 1), the response X in the telephone receiver depends

upon B, that is, upon the extent to which the incoming wave is modulated. In order to obviate any error due to this effect it is necessary to make $B=1$ by completely modulating the test wave, which can then be specified by the expression

$$A \sin pt\,(1-\sin qt) \tag{17}$$

It is to be remembered that a modulating device for which the power series (3) does not converge sufficiently rapidly to enable us to neglect terms of higher power than the second, will not produce a modulated wave as specified by (17). The characteristic of the thermionic vacuum tube, however, converges so rapidly that it satisfies (17) within the limits of experimental error. The wave used in the experiments described in this paper was a 300,000 cycle wave completely modulated by 800 cycles with a thermionic tube and can therefore be regarded as being characterized by expression (17). Complete modulation was secured by predetermining the negative grid voltage necessary to reduce the current to zero for the plate voltage used and then making the peak value of the audio frequency input voltage equal to $\frac{E_B}{\mu}-E_c$, where E_c is the constant grid battery voltage.

6. Function of the Receiver Shunt. The receiver shunt furnishes a quick and simple means of measuring detecting currents. Its series and shunt resistances each consist of a number of separate non-inductive resistance units, so arranged that definite pairs are connected in circuit for each adjustment of the shunt, their values being so chosen that the total impedance into which the generator U works remains constant. The value of a constant impedance shunt becomes apparent when considering the measurements in the determination of the detecting current as a function of the input signal strength.

For convenience in representing the measurements, the relation between the generator current i_1, as measured by the ammeter A, and the detecting current i can be represented by the well known equation for current attenuation by a line or cable:

$$\frac{i_1}{i}=e^{ad} \tag{18}$$

where e is the base of the natural logarithms, d is the length of cable of which a is the attenuation constant per unit length. The constant a is, of course, purely arbitrary. In conformity with practice among telephone engineers we shall make a equal to the attenuation constant of the so-called "standard number

19 gauge cable," namely, 0.109 per mile at a frequency of 800 cycles per second.* This reference cable has a capacity of 0.054 microfarad and a resistance of 88 ohms per mile at a frequency of 800 cycles per second. The current attenuation can then be expressed in miles, d, of the chosen standard cable of reference, the receiver shunt being calibrated in terms of two-mile steps. Length of cable forms a very convenient unit of measurement in cases where the telephone receiver is used as the measuring instrument. This unit has long been in use in telephone practice and I would urge its general adoption in measurements relating to radio detectors. The fact that the current ratio $\frac{i_1}{i}$ changes rapidly with the cable length d is not a disadvantage attending the use of this unit because small changes in current are not easily detected with a telephone receiver, which, on the other hand, gives sufficiently accurate readings since it is the instrument used in practice for giving us sense impressions of current. In fact, it is usually sufficient if the receiver shunt is calibrated in steps of two miles of standard cable. With a little practice such a calibration allows of an estimate to within one mile. It is to be understood that the relation between the length d of cable and the current ratio depends upon the chosen value of a which must be agreed upon. Wherever reference is made in the following to miles of cable the value of a will be understood to be 0.109. For convenience of reference the relation between d and $\frac{i_1}{i}$ for this reference cable is given in Table 1.

TABLE 1

Miles of Standard Cable d	Current Ratio $\frac{i_1}{i}$	Miles of Standard Cable d	Current Ratio $\frac{i_1}{i}$
5	1.72	50	232
10	2.97	60	689
15	5.13	70	2.05×10^3
20	8.85	80	6.08×10^3
30	26.3	90	1.82×10^4
40	78	100	5.35×10^4

* 1 mile = 1.6 km.

7. Experimental Results. Since the detecting current i and input signal voltage e are connected by

$$i = a\,e^2 \tag{4}$$

we get by substitution in (18):

$$d = -2\,K\,log_{10}\,e + C$$

where

$$\left.\begin{aligned} C &= K\,log_{10}\frac{i_1}{a} \\ K &= \frac{2.3026}{a} = 21.13 \end{aligned}\right\} \tag{20}$$

The detection coefficient a can now be obtained in a simple way by successively applying different input voltages e to the tube and every time adjusting the receiver shunt S (Figure 1), so as to make the note in the receiver T of equal intensity for both positions of the switch W. The intercept C ($log\,e = 0$) of the straight line (19) gives a in terms of the known values of K and i_1:

$$log\,a = log\,i_1 - \frac{C}{K} \tag{21}$$

The source of audio frequency which supplied the auxiliary note was used also to modulate the radio frequency. The circuit arrangement is shown in Figure 5. U represents the source of audio frequency current. What was actually used for this purpose in these experiments was a vacuum tube oscillator. The microphone generator shown in the diagram serves the purpose as well and has been used in portable vacuum tube testing sets. Its principal of operation is the same as that of an interrupter altho it is much superior, the interrupter being, on account of its unreliability and need of constant adjustment, practically useless for accurate measurements. The microphone generator which has long been used in telephone testing work, has the advantage that its circuit is never broken as in the case of the interrupter, the resistance of the carbon being merely varied harmonically by the effect of the flux in the coil on the diafram. It is therefore free from the usual troubles attending the use of an interrupter, such as sparking and consequent corroding of the contacts. It operates on about 3 to 5 volts d.c. The a.c. obtained from it is transmitted thru a filter F to give a pure note of 800 cycles. This and the radio frequency oscillations obtained from the vacuum tube oscillator O are both impressed on the input of the modulator M, the

resulting modulated wave being impressed on the detector D. By means of the switch, W, the low frequency current can be transmitted either to the modulator or, thru the shunt S, to the telephone receiver T. Since the receiver had an impedance of

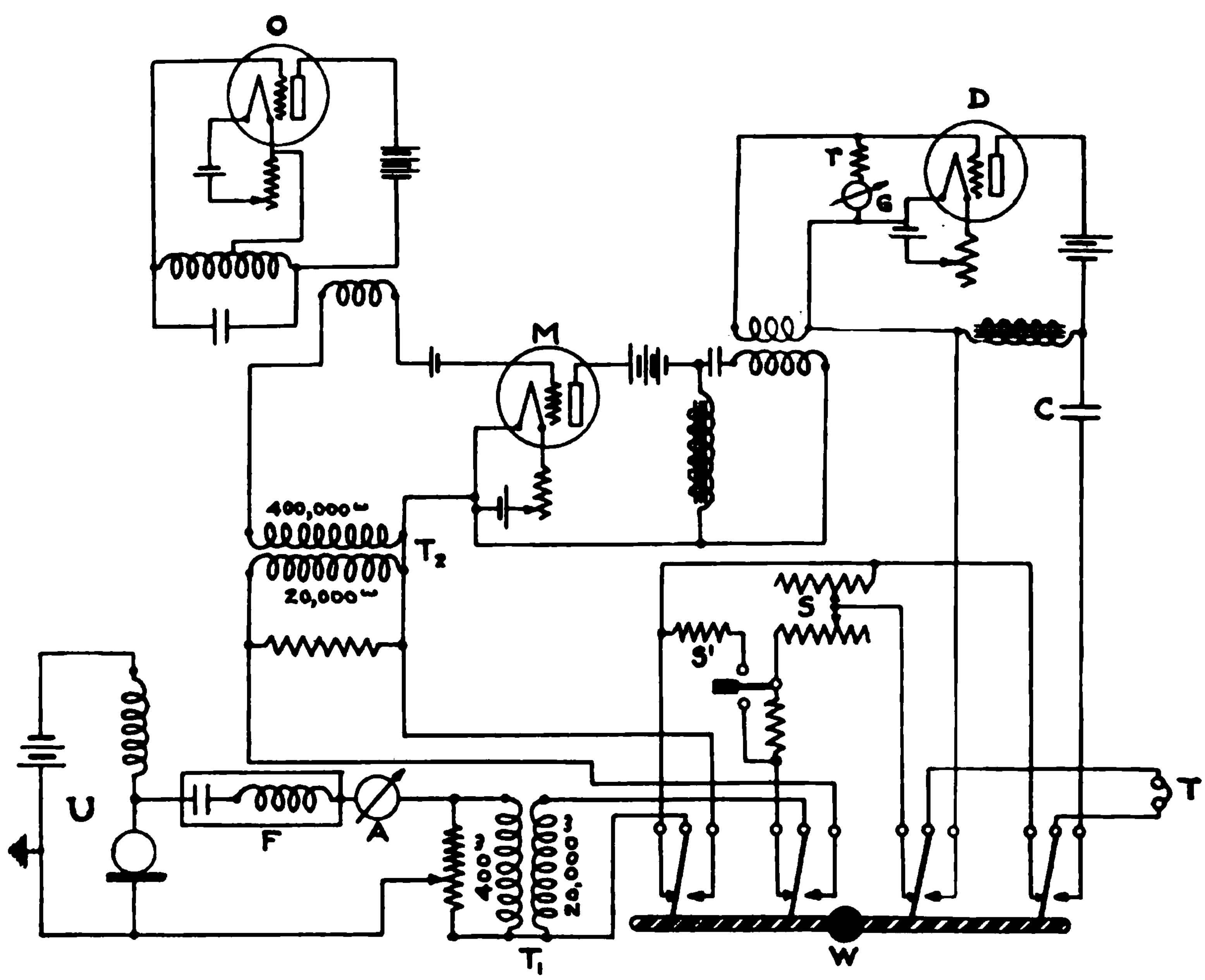

FIGURE 5

20,000 ohms the transformers T_1 and T_2 were inserted to insure that the generator U works into the same impedance for both positions of the switch W. If the audio frequency current is measured by A on the low side of the line a small correction might be needed for loss in the transformer T_1. However the audio frequency vacuum tube oscillator used in these experiments gave sufficient power to enable the current to be measured directly in the 20,000-ohm line. The receiver shunt S gave a maximum attenuation of 30 miles of standard cable $\left(\frac{i_1}{i} = 26.3\right)$. For higher attenuations extra shunts S', each of 30 miles, could be added.

To insure complete modulation of the wave impressed on the detector the modulator M can be made to measure its own input voltage according to the principle of the vacuum tube

voltmeter of R. A. Heising[5]. If E_B and μ are the plate-filament voltage and amplification constant of the modulator, the negative grid voltage necessary to reduce the current in the plate circuit to zero is $\frac{E_B}{\mu}$. Complete modulation can then be obtained by applying a constant negative grid voltage $E_c = \frac{E_B}{2\mu}$ and making the peak value of the audio frequency input voltage e_1 also equal to $\frac{E_B}{2\mu}$. This can be done by first adjusting the negative grid voltage to a value $\frac{E_B}{\mu} + \frac{E_B}{2\mu}$ and then applying the low frequency a. c. and adjusting it until a current meter in the output circuit of M just begins to show a deflection. The peak value of the a. c. input is then equal to $\frac{E_B}{2\mu}$. The radio frequency can be measured in the same way and should be weaker than the audio frequency. For operating the modulator E_c must of course be finally adjusted to its proper value.

The accuracy with which the linear relation given by equation (19) holds is shown in Figure 6 in which the crosses and circles represent observations made by two different observers on different days. The close agreement of the slope, 42.2, of

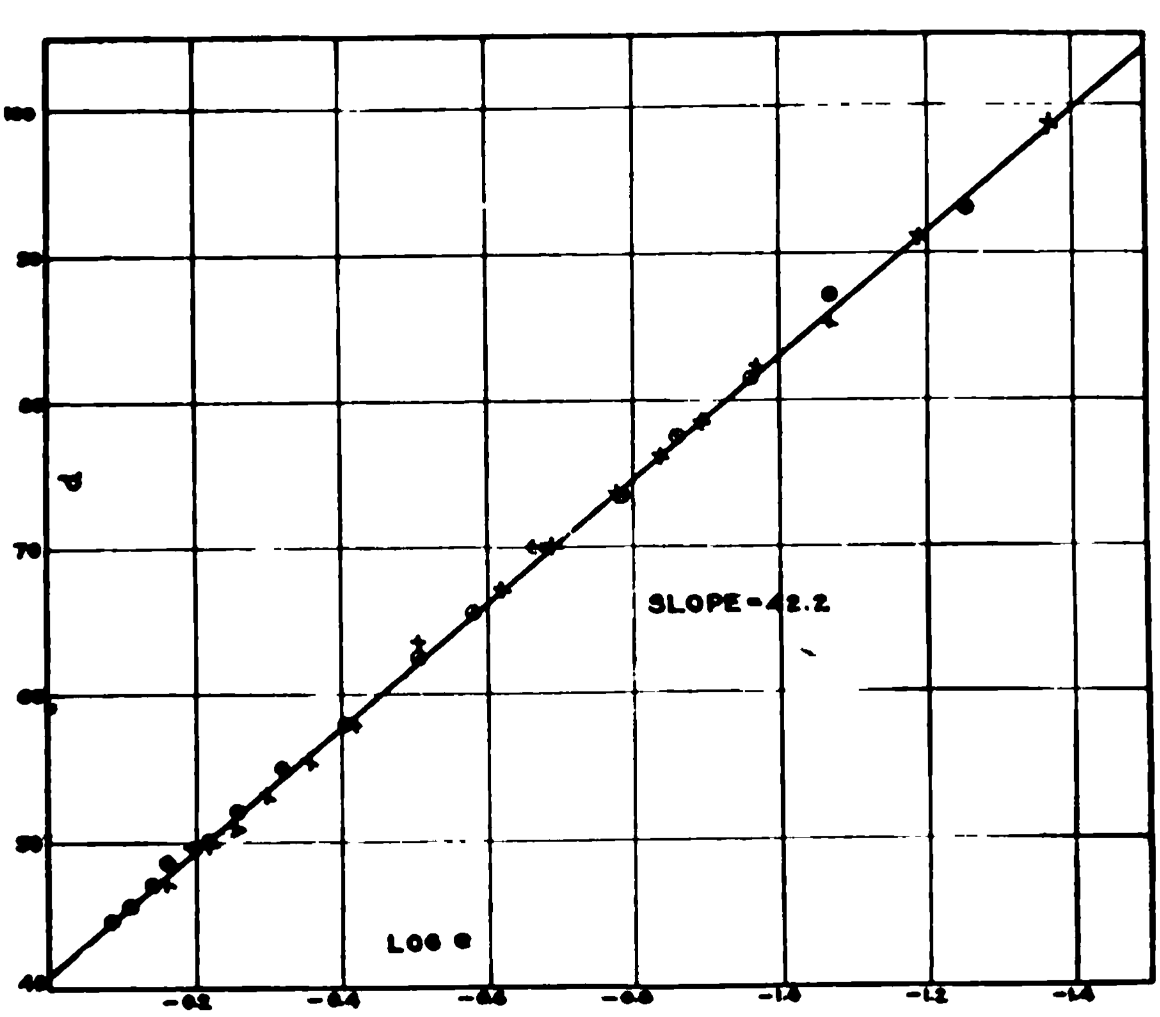

FIGURE 6

[5] R. A. Heising, U. S. Patent 1,232,919.

this line with the theoretical value, $2K=42.26$, verifies equation (4). In measurements of this kind, where the intensities of two notes are compared, the influence of extraneous noises is small, almost negligible compared with their influence on measurements by the audibility method. The error in the observations was due largely to the fact that the detecting current contains a weak harmonic component (see equation 14) which to some observers seems to have a somewhat disturbing influence. If necessary this harmonic could easily be filtered out by the insertion of an appropriate inductance in series with the condenser C in the output of the detector.

The current i_1 as measured by the meter A, inserted in the 20,000-ohm line, was 3.10×10^{-3} ampere and the intercept for $e=1$ of the line shown in Figure 6 is 40.8. Hence the detection coefficient $a=36.2\times10^{-6}$ amp./(volt)^2.

Table 2 gives the results obtained by a number of different observers, the observations being made on three successive days.

TABLE 2

Observer	Slope $2K$	Intercept C	Detection Coefficient Amp./(volt)^2
A	42.6	39.9	41.6×10^{-6}
A	42.6	40.0	41.0×10^{-6}
A	41.0	41.1	31.0×10^{-6}
B	42.7	39.7	42.8×10^{-6}
B	42.0	41.4	33.5×10^{-6}
B	42.1	41.4	33.7×10^{-6}
C	42.2	41.7	32.9×10^{-6}
D	41.4	41.4	31.0×10^{-6}
E	41.6	41.0	33.1×10^{-6}
Mean	42.0	40.8	35.6×10^{-6}

The input signal voltages concerned in these measurements are the root-mean-square values of the completely modulated wave. If it is necessary to express them in peak values they must be multiplied by $4/\sqrt{3}$ instead of by $\sqrt{2}$ as in the case of an unmodulated wave. This can be seen by making B equal to unity in equation (16) and noting that the peak value of the wave is $2A$ (equation 12).

8. Detecting Efficiency. The detection coefficient a, which gives a measure of the audible component of the current in the output of the detector, is not sufficient for expressing the figure of merit of the device, because it depends on the type of telephone receiver used. The impedance of the receiver must be so chosen as to give maximum response and the effect expressed in terms of the power dissipated in the receiver, which must, of course, be computed from its resistance.

When the thermionic detector is used without a blocking condenser in the grid circuit, as was the case in these experiments, the most efficient operation is obtained when the grid is kept sufficiently negative so as not to take current. The power consumption in the input circuit is then practically zero. The response in the output circuit depends on the potential difference established between filament and grid and not necessarily upon the power consumption in the input circuit. The detecting efficiency can, therefore, not be expressed as the ratio of audio frequency output power to radio frequency input power. The same reasoning applies to a consideration of the power loss in the input coil and condenser when using the customary input resonant circuit obtained by replacing the resistance and galvanometer G (Figure 1) by a condenser. Obviously if the power loss is a minimum the input voltage, and therefore also the detecting current, is a maximum.

This consideration furnishes an answer to the question as to whether the audion detector can amplify, that is, can give more audio frequency power in the output than radio frequency power expended in the input. Since the tube is a potential operated device the answer to this question must be in the affirmative. All that is necessary to verify this experimentally, is to impress a convenient voltage on the input of the tube, let us say by an arrangement such as shown in Figure 1. Practically all the input power consumption takes place in the resistance r and can be made as small as we please by making r sufficiently large and adjusting the input coupling so as to keep the input voltage constant. In this respect the thermionic detector differs from other types of detectors such as the Fleming valve and crystal detectors. Detection by these devices depends upon rectification of the incoming current. They can, therefore, never give more audio frequency power than the radio frequency power consumed in the input, in fact, they can never give as much. The audion, on the other hand, at least when operated without a grid condenser, does not detect by virtue of rectification of the incoming current.

The incoming current is not rectified at all and detection takes place solely by virtue of the curvature of the plate current characteristic, which makes possible an unsymmetrical release of energy supplied by the plate battery.

In view of these considerations it is best to express the detecting efficiency in terms of the relation of output audio frequency power to input radio frequency voltage. It is, therefore, given by

$$\delta = a^2 t \tag{22}$$

where a is the detection coefficient and t the resistance of the telephone receiver.

The receiver used in these experiments had an impedance of 20,000 ohms and a resistance of 6,400 ohms at 800 cycles per second. Hence, taking the mean value of a given in Table 2, the detecting efficiency is

$$\delta = 8.1 \times 10^{-6} \text{ watt}/(\text{volt})^4$$

The tube on which these measurements were made was one of the Western Electric type, V. T. 1 (Figure 7-a), that was specially designed for airplane radio telephone service and used by the United States Signal Corps for this purpose. This type

Figure 7a

Figure 7b

of tube does not give as high an efficiency as some of the pre-war tubes, but it was designed to operate on low power consumption, its operating plate voltage being about 20 volts and the power consumed in heating its filament 2.2 to 3.5 watts.

Figure 7-b shows a pre-war type of Western Electric tube which has been in use in this laboratory since the early part of 1916. While this tube has a higher detecting efficiency than that shown in Figure 7-a, the power consumed in its filament is about twice as much and it operates on a plate battery of 30 volts. Another type of pre-war tube is shown in Figure 7-c. This tube has an amplification constant of 40, and can be used as a voltage amplifier or detector.

FIGURE 7c

FIGURE 7d

Smaller types of detector and amplifier tubes have since been developed which require only a few tenths of a watt for heating the filament. Such a tube is shown in Figure 7-d. With a filament voltage of 1 volt, filament current of 0.18 ampere and a plate voltage of 10 volts, this tube has a detecting efficiency of 4.3×10^{-6} watt/(volt)^4. The filament can be operated on one dry cell, its operating voltage ranging from 1.0 to 1.5 volts.

The detecting efficiency of this tube is therefore about one-half that of the V. T. 1 tube. The actual effect on the ear due to this difference is not large. A ratio of two in the power corresponds to a difference of about three standard cable miles. A difference of one standard cable mile is hardly noticeable unless the comparison be made directly. For this reason also the detection coefficients given in Table 2 should be regarded as showing very small variations. The more correct measure, as far as the ear is concerned, is the detection coefficient expressed on the logarithmic scale, that is, the values of the intercept C' given in the third column of Table 2, and which are expressed in miles of standard cable.

An idea can be obtained regarding the intensity of the sound produced when power of the order given by these tubes is dissipated in the receiver, by noting that the power dissipation necessary in this receiver to give the least audible signal is about 3×10^{-12} watt. From the above value of the detecting coefficient the input voltage necessary to give the least audible signal is therefore about 0.025 volt.

9. Comparison of Detectors. The constancy with which vacuum tubes can now be made to operate as detectors brings the testing of detector tubes by means of comparison circuits within commercial possibility. Comparison methods have a practical advantage in that they do not necessitate such accurate calibration of the quantities involved. Once the detecting efficiency of a tube has been determined the tube can be used as a standard of comparison to obtain the efficiency of any other tube by the simple circuit shown in Figure 8. The input voltage

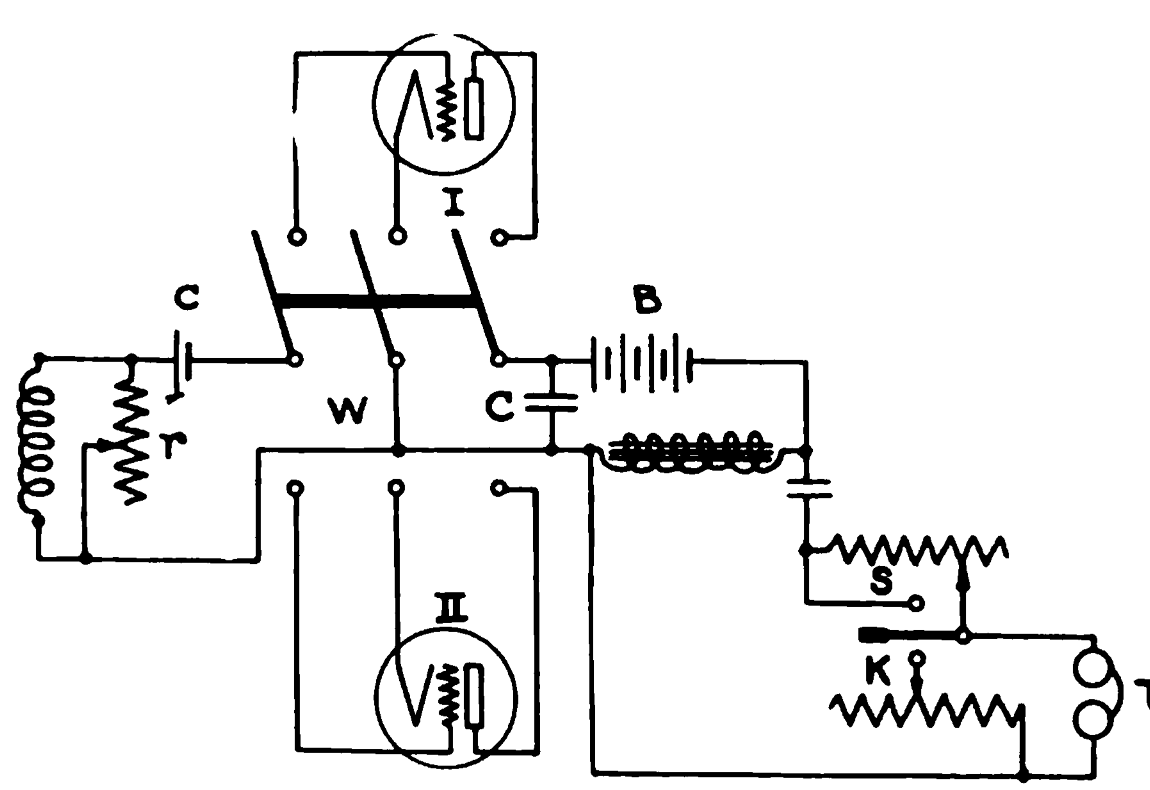

Figure 8

can be adjusted to any desired value by adjusting r, and need not be known accurately in this case, it being sufficient to know that it lies within the range of voltages encountered in practice. By means of the switch W either detector can be inserted in the circuit and the receiver shunt S adjusted until the note in the receiver T is of equal intensity for both positions of the switch W.

C is a radio frequency leak. By means of the key K the shunt can be thrown into or out of circuit according as the switch W connects the tube of higher efficiency or the other. If i_1 and i_2 be the detecting currents obtained from the tubes 1 and 2, and a_1 and a_2 their detection coefficients, then since the input voltage is the same for both

$$\log\frac{i_1}{i_2}=\log\frac{a_1}{a_2}=\frac{d_1-d_2}{K}=\frac{d}{K}$$

where d is the adjustment of the shunt in units depending on the units of K. As before we can express d in miles of standard cable by making $K=21.13$. The detecting efficiency can then be obtained from

$$\delta_2=\left(\frac{a_2}{a_1}\right)^2\delta_1 \tag{23}$$

10. Modified Audibility Method. The audibility or "shunted telephone" method has been frequently applied in attempts to measure the strength of received signals in long distance radio communication and has also been used in obtaining an idea of the sensitiveness of detectors. This method consists in shunting the telephone receiver connected in the output of the detector and determining the value of the shunt necessary to reduce the current in the receiver to such an extent that dot and dash signals can just barely be differentiated. The ratio of the total current in the receiver and shunt, that is, the detecting current, to the current in the receiver alone, measures the "audibility" and can be computed from the resistance of the shunt and the impedance of the telephone receiver. When we consider the possibility of accurate measurements, this method is open to serious objections. In the first place it is liable to considerable error because the measurement of least audible signals is made difficult by the influence of extraneous noises, such as room noises and static. It furthermore depends to an appreciable extent upon the condition of the observer, so that the current necessary to give least audible signals will vary from time to time even with the same observer. These disad-

vantages make this method quite unreliable for purposes of determining detecting efficiency.

Secondly, the way in which the audibility method is ordinarily used does not make provision for the change in effective impedance when the shunt resistance is varied. This would give misleading results, since the detecting current, that is, the audible component of the current in the output circuit of the detector tube, depends upon the relative values of the internal output impedance of the tube and the impedance into which the tube works. It is therefore necessary in all measurements of this kind to adjust these impedances properly and keep them constant thruout the measurements. If the audibility method is to be used the "audibility box" should be so designed that any variation in the shunt resistance is accompanied by an addition or subtraction of an equivalent resistance so as to keep the total impedance of the circuit constant. This could be done with the scheme that will now be described.

The audibility measurements given here were made with the object of determining the possible error to which measurements would be subject when the audibility method is used, for example, for the determination of the strength of incoming signals. The fact that the current necessary to give least audible signals has different values for different observers, and is therefore incapable of objective determination does not of itself rule out the audibility method for the measurement of signal strength, since the detector set could first be calibrated by determining the audibility for known input signals and then used by the same observer to make the final measurements. Hence, assuming that extraneous noises could be effectively cut out, the possibility of adapting this method to such measurements would depend upon the extent to which the observer's conception of least audible signal remains constant during the time that elapses between his calibration of the set and the making of his final measurements. It is hardly necessary to say that the whole set must remain unchanged, especially the tube and the telephone receiver.

The circuit by which the variability of least audible signal was studied is shown in Figure 9. This circuit allows of audibility measurements under constant circuit conditions. The tube used was the same one on which the above measurements were made. It was a Western Electric "standard" the detecting efficiency of which had not changed to any noticeable extent in the course of seven months during which time it was in fre-

quent use. Any variability of the least audible signal as determined by this set was due entirely to the observers.

The input signal wave was, as before, a completely modulated wave and could be measured with the Duddell galvanometer G. In order to keep the impedance of the output circuit constant the receiver shunt described above was used to shunt down the current in the telephone receiver. The choke coil, L, was inserted

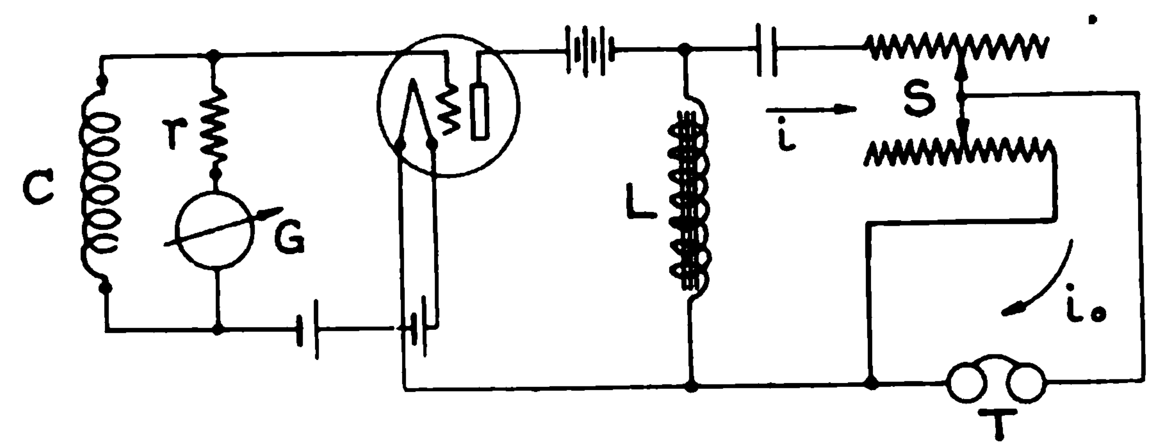

FIGURE 9

to keep the impedance of the tube constant. This was necessary because, altho the *impedance* of the circuit remains constant for all adjustments of the shunt, its d. c. *resistance* varies and if the coil L were omitted the variation in resistance would result in a variation of the d. c. potential difference between plate and filament with a consequent variation in the tube impedance.

The use of the receiver shunt makes it possible to express the "audibility" in a simple way. If i be the detecting current and i_o the least audible signal current in the receiver the audibility $\frac{i}{i_o}$ can be expressed in miles d of cable by equation (19) where the intercept C $(e=1)$ is now given by:

$$C = K\,(log\,a - log\,i_o) \tag{24}$$

and gives a measure of the audibility efficiency in miles/(volt).[2]

The simple linear relation (19) makes it possible to obtain the audibility efficiency as the average of a large number of observations. A number of such observations plotted against the logarithm of the corresponding input voltages are shown in Figure 10. Altho the individual points vary considerably from the average they are nevertheless grouped evenly about the straight line, the slope of which, namely, 41, is quite close to the theoretical value, 42.26. In no case was there any difficulty in verifying the linear relation.

The slope of the line depends only upon the attenuation constant of the shunt and the simple quadratic relation (4).

The intercept, on the other hand, is influenced also by the detection coefficient, a, of the tube itself and by what constitutes the least audible signal current, i_o, for the particular telephone receiver used and for the observer at the particular time of making the measurements. The detection coefficient and the

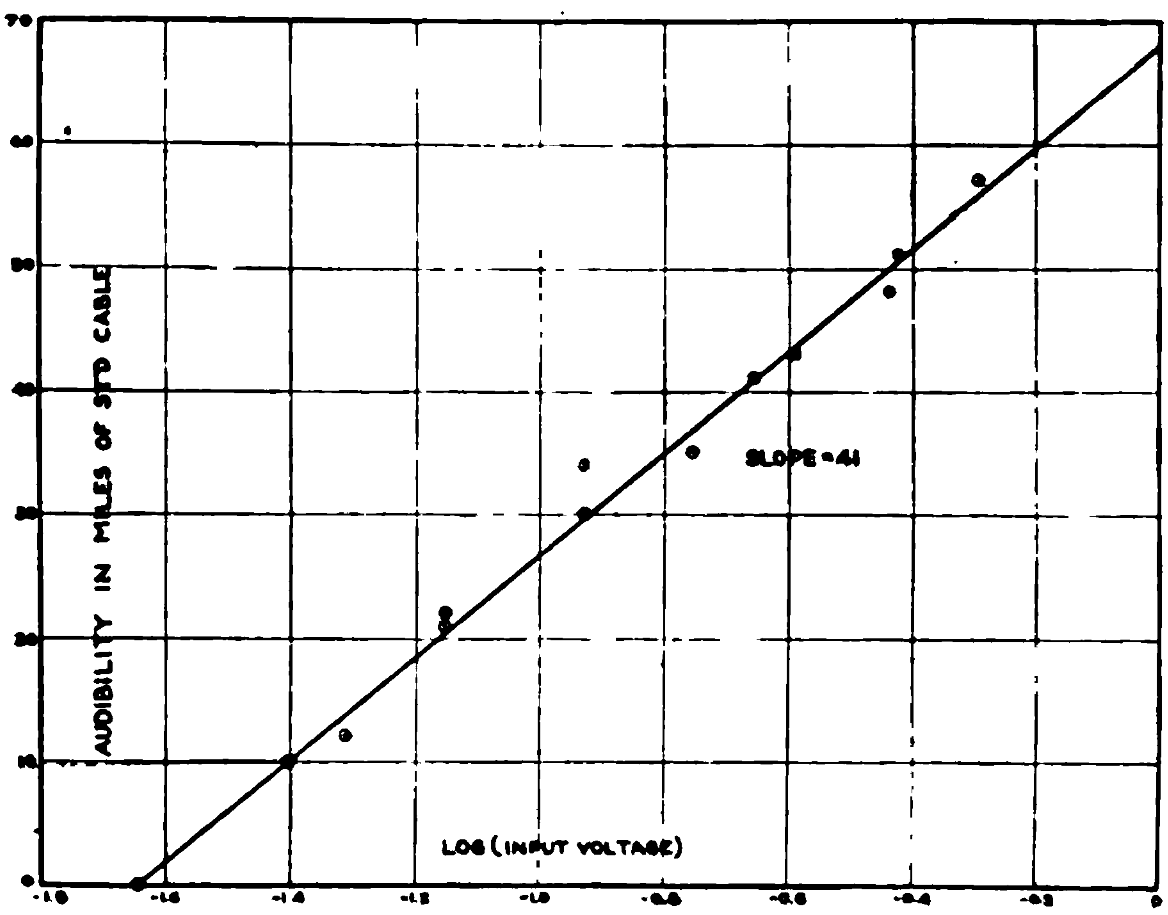

FIGURE 10

attenuation factor K certainly do not vary enough to be noticed. Any variation in the intercept C will, therefore, be due to the least audible current. Hence a number of separate determinations of C could be used to give an indication of the accuracy of the audibility method. Now, the least audible current depends firstly upon sensitiveness of the telephone receiver. By using the same receiver thruout the measurement any variation due to this factor can be regarded as negligibly small compared with the much more disturbing variation in conception of least audible signal on the part of the observers. Since this conception varies with different observers as well as with the same observer at different times, measurements were performed by four different observers over a period of eight days. The results are shown in Table 3 and were obtained from curves such as shown in Figure 10. They therefore give the average of over 350 observations. The fourth column gives the audibility efficiency in miles/(volt)^2, ($\log e = 0$). The corresponding audibilities expressed in current ratio, determined from Table 1, are given in the adjacent column. The last column gives the input voltage (r.m.s.) necessary to give a detecting current equal to the least audible signal current ($d = 0$).

TABLE 3

Observer	Time	Slope 2 K	Audibility Efficiency, C Miles/(Volt)^2	Audibility Efficiency, C $\frac{i}{i_0}$/(Volt)^2	Input for Least Audible Current (Volts)
A	1st day A.M.	42.7	60.9	760	.037
	2nd " A.M.	41.4	61.6	820	.036
	2nd " P.M.	45.0	64.6	1140	.038
	3rd " A.M.	44.2	62.8	940	.030
	5th " A.M.	39.5	60.1	700	.033
	5th " P.M.	44.6	65.7	1280	.028
	6th " A.M.	43.7	68.6	1720	.025
	6th " P.M.	43.1	68.9	1800	.022
	7th " A.M.	40.8	67.6	1550	.023
	8th " A.M.	42.3	69.8	1980	....
	Mean:	42.7	65.0	1260	.030
B	2nd day A.M.	40.4	62.4	900	.028
	2nd " P.M.	43.0	65.8	1300	.030
	3rd " A.M.	38.0	61.6	820	.024
	5th " A.M.	41.5	64.1	1080	.028
	5th " P.M.	41.1	63.7	1040	.028
	6th " A.M.	42.1	64.0	1070	.030
	6th " P.M.	42.4	65.5	1250	.028
	7th " P.M.	43.2	68.1	1640	.026
	8th " A.M.	41.4	64.5	1130	.028
	Mean:	41.5	64.4	1130	.028
C	3rd day A.M.	43.6	66.6	1400	.024
	5th " A.M.	40.8	65.6	1260	.025
	5th " P.M.	43.3	65.4	1240	.031
	6th " P.M.	46.0	66.2	1340	.037
	7th " P.M.	41.2	67.8	1580	.023
	8th " A.M.	43.0	66.5	1380	.029
	Mean:	43.0	66.4	1360	.028
D	7th day A.M.	44.0	77.2	4840	.018
	7th " P.M.	44.0	73.5	3140	.021
	Mean:	44.0	75.4	4000	.020

It will be seen from these observations that the audibility varies quite considerably with different observers, and even with the same observer at different times. It must be remarked that these measurements were not performed under ideal conditions as regards freedom from noise, outside noises from the city streets being sometimes distinctly noticeable. Ideal conditions are, however, not practicable and the value of such measurements could reasonably be expected to be determined to a great extent by the accuracy with which they can be made under normal conditions. It is possible that the wide discrepancy in the reported error attending audibility measurements

[6] See L. W. Austin, PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS August, 1917, page 239, and the discussion by J. Mauradian, page 248.

(the error ranging from about 20 per cent to several hundred per cent)[6] is due to the varying noise conditions under which the measurements were performed.

11. APPLICATION TO DETERMINATION OF SIGNAL STRENGTH. When considering the practical problem of measuring the strength of received signals, the audibility method has the one advantage that it does not require an auxiliary source of audio frequency. Its inaccuracy, however, would hardly recommend it for such measurements. The method described in the first part of this paper has the advantage of greater accuracy, especially when considering the influence of static disturbances. The constancy of the tube makes it possible to determine the efficiency of the set beforehand. From the observed detecting current the received input voltage can then be obtained, from which the received power can be calculated when the constants of the input circuit are known. For accurate measurements it would, of course, be necessary to consider the form of the incoming wave. When measuring weak signals received on long distance communication, the heterodyne method could be used. By putting the detecting current $i = b\, e\, e'$, instead of equation (4), where e is the signal voltage and e' the local input voltage, a simple linear relation can be obtained similar to equation (19). The local voltage e' could be made sufficiently large to enable it to be measured with the detector itself in the manner explained above (section 7) in connection with the measurement of the voltages impressed on the input of the modulator.

12. PRACTICAL APPLICATIONS. When it is necessary to test a large number of detector tubes, the circuit shown in Figure 5 is somewhat impracticable, because such a circuit requires careful adjustment of the operating parameters, such as the input radio frequency voltage impressed on the detector, the audio frequency current delivered by the generator U, and so on. The procedure that was adopted and found more practicable was to use this circuit to standardize, by means of the method explained above, carefully selected tubes the detecting efficiency of which was found to remain constant. These tubes were then used as comparison standards to test the factory product by means of a circuit like that shown in Figure 8. The tubes could then be compared with the standard under the actual circuit conditions under which they may be required to operate in practice. Figure 11 shows a comparison test circuit designed to test detectors which are intended for use in conjunction with

an amplifier. The circuit is arranged in a box, shown in Figure 12, with terminals for connecting the batteries and jacks for measuring the filament and plate voltages. It was necessary to secure a simple means of producing constant modulated radio frequency oscillations to serve as the source of input voltage. Ordinarily this would require a radio frequency oscillator, an audio frequency oscillator and a modulator. To avoid

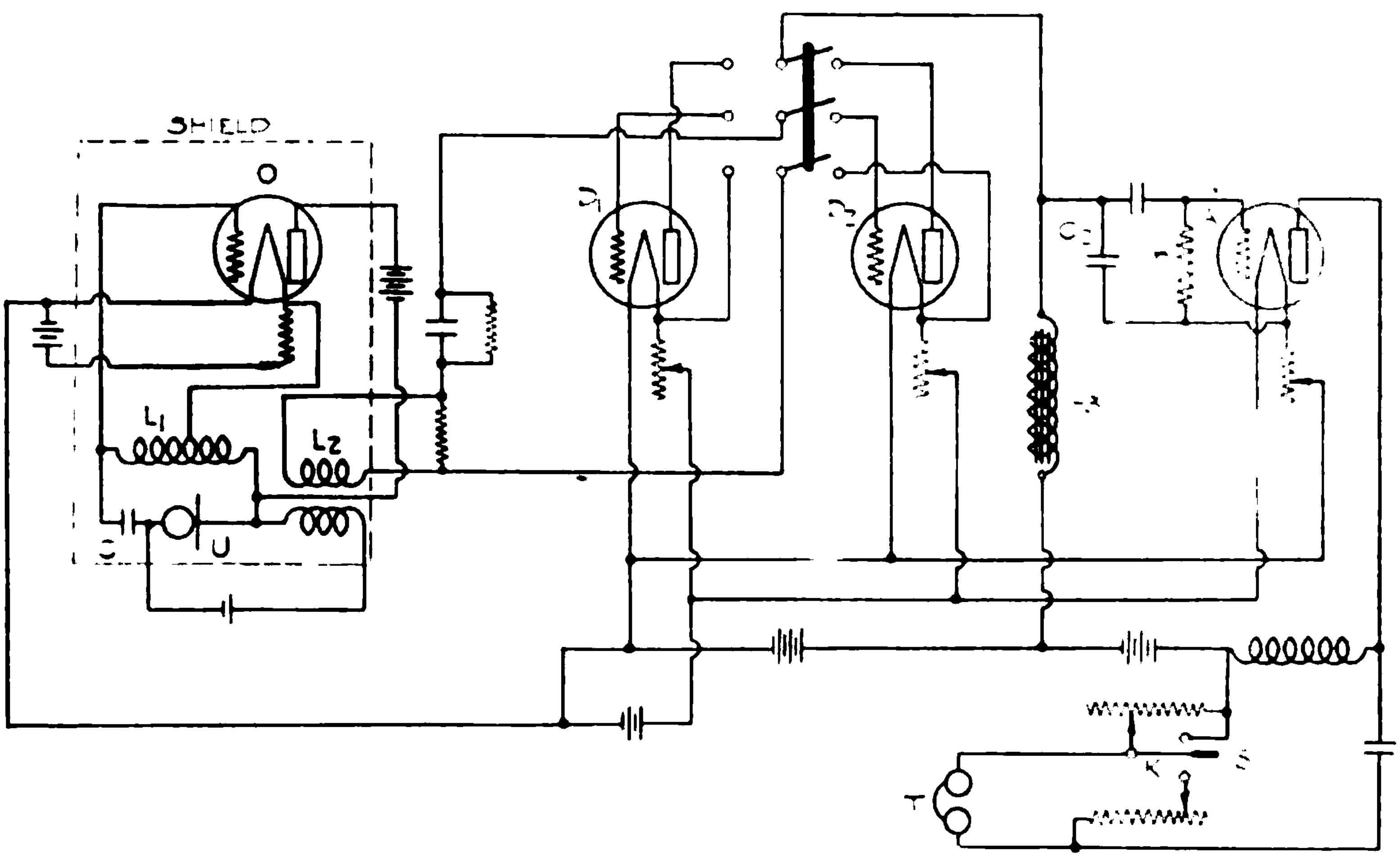

FIGURE 11

using such a cumbersome and rather impracticable arrangement the oscillator tube, *O* (Figure 11) was made to produce modulated high frequency oscillations directly with the aid of a microphone generator, *U*. The carbon button of this generator was inserted directly in the oscillating circuit, $L_1\,C_1$, as shown in the diagram. The microphone causes an audio frequency variation of the resistance of the oscillating circuit and thus serves to modulate the radio frequency oscillations. The modulation produced by this means is not complete, but complete modulation is not necessary when *comparing* detectors. The audio frequency component in the output of the detector is amplified by the tube, *A*. C_2 is a high frequency leak and *r* a high resistance acting as a grid leak to prevent the grid from building up a charge. The difference in detecting efficiency of the two tubes can be measured by means of the shunt, *S*, in the manner explained in section 9, and can, as shown in the measurements described above, be most conveniently expressed

in the logarithmic scale. The shunt used in these test sets had a constant $K=21.13$, thus giving the difference in detecting efficiency in miles of standard number 19 gauge cable.

FIGURE 12

APPENDIX

At the beginning of this paper mention was made of a simple method of measuring the amplification of audion tubes. Altho there is nothing new in this method which has been in use in this laboratory for many years, it is not generally known and may, therefore, be briefly described here. The circuit arrangement whereby the amplification can be measured is shown in Figure 13. Audio frequency currents are supplied by the generator, U, and are passed thru the filter to obtain a pure note of, say, 800 cycles per second. By operating the switch, W, this current can be transmitted either directly to the telephone receiver, T, or thru the tube and a receiver shunt, S. By adjusting the shunt until the note heard in the receiver is of the same intensity for both positions of the switch, W, the current amplification given by the tube is equal to the attenuation produced by the shunt, and is given by equation (18), or

$$d = K \log \frac{i_2}{i_1}$$

where i is the current in the receiver and i_2 the current before being attenuated by the shunt. The reading, d, of the shunt

usually gives the amplification in miles of standard cable. The actual current ratio can, as before, be obtained from Table 1. The power amplification, which is a more desirable expression, and means more than current amplification, can be obtained from the above equation by dividing the factor K by 2.

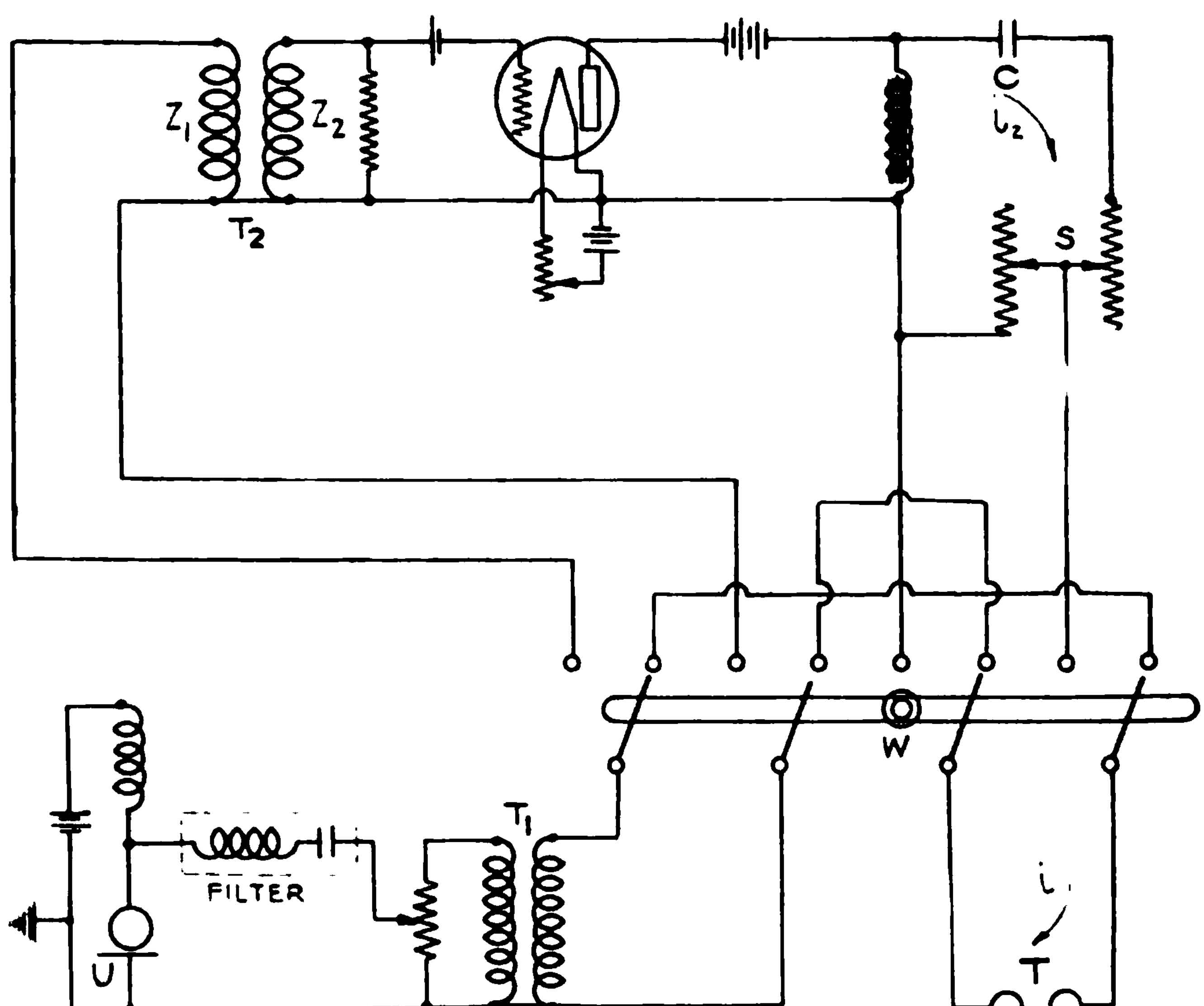

FIGURE 13

The generator, U, can be of any type that gives constant audio frequency current. In Figure 13 it takes the form of a microphone generator. On account of its compactness, this is a desirable source of audio frequency currents, when the test set is to be small, compact, and portable. A portable Western Electric test set (made for the United States Signal Corps) is shown in Figure 14.

Care must be taken, when using this method that the current in the secondary of the transformer, T_1, is the same when connected to the input of the tube as when connected to the telephone receiver. For this purpose the impedance, Z_1, of the transformer, T_2, is made equal to the impedance of the receiver. The secondary of transformer, T_2, is wound to have as high an impedance as possible, to impress the maximum possible voltage on the grid circuit of the tube. When, as in

FIGURE 14

this case, the receiver has an impedance equal to the plate resistance of the tube, it can be connected directly to the output circuit of the tube, as shown in the diagram, the receiver shunt being so designed that when placed between the tube and the receiver, the total impedance of the output circuit remains constant for all adjustments of the shunt. If the impedance of the receiver is very different from the plate resistance a transformer of the proper impedance ratio must be inserted in the output circuit.

In conclusion I wish to acknowledge the assistance of Mr. R. H. Wilson in carrying out the experiments described above.

Research Laboratory of the
American Telephone and Telegraph Company and
Western Electric Company, Incorporated,
New York City.

March, 1919.

SUMMARY: There is described a method of determining the detecting efficiency of vacuum tubes by feeding them with radio frequency current modulated at audio frequency. The theory of the method is given.

In connection with such measurement, the use of a receiver shunt calibrated in terms of miles of "standard number 19 gauge cable" is advocated.

For producing the radio frequency current for such measurement, a tube oscillator was employed. The audio frequency modulation was obtained either by a second tube or by a microphone howler. The apparatus is described in detail, and illustrated. Experimental data are given.

A method for comparing the detecting efficiencies of a standard tube and of a tube under test is also described, and the apparatus is shown.

In an Appendix, a simple method of measuring tube amplification is [illegible]d

DISCUSSION

John M. Miller: In connection with the method which has just been described for measuring the detecting efficiency of a three-electrode vacuum tube, it may be of interest to outline the method which has been independently developed at the Bureau of Standards for making these measurements. The general requirements of such a method are somewhat obvious, that is, a relation must be established between the radio frequency input to the detector and the audio frequency output. In order to obtain a source of completely modulated radio frequency for calibrating detector tubes, we have made use of the arrangement shown on the left of Figure 1. In the plate cir-

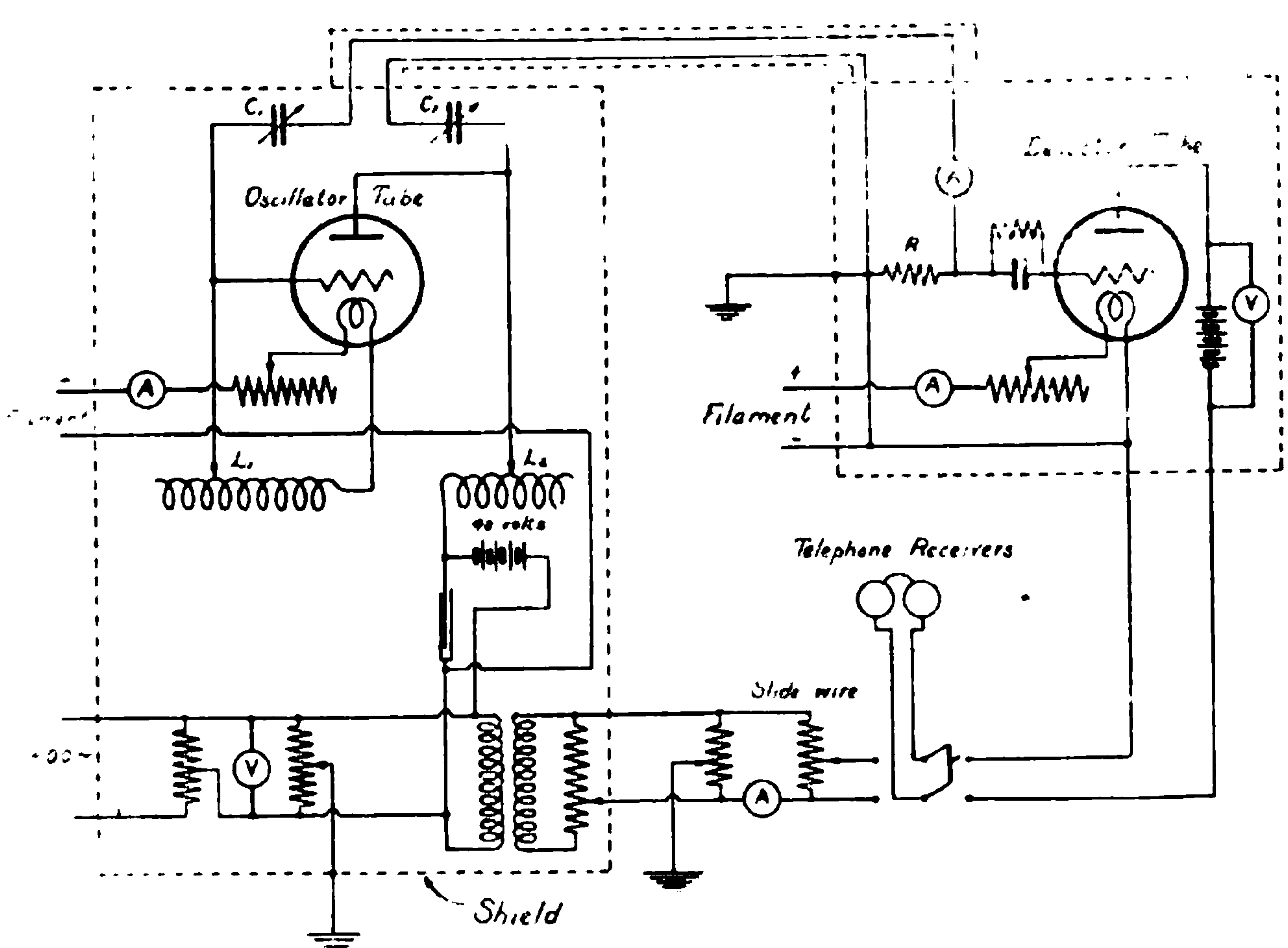

FIGURE 1

cuit of the oscillator tube which supplies the modulated radio frequency emf. to the detector tube there is an audio frequency voltage furnished by an alternator which is in series with the customary d. c. battery. The peak value of the alternating voltage is adjusted to equality with the d. c. voltage so that the total plate voltage varies from zero to a maximum value of twice that of the d. c. battery. A calibration of the output of the oscillator tube with respect to the plate voltage shows a linear relationship between output current and plate voltage thruout the range of voltage used, so that under the action of

the combined a. c. and d. c. voltages the output current will be completely modulated with a sinusoidal envelope. A known emf. of this modulated radio frequency is supplied to the input of the detector tube, by measuring with a vacuum thermocouple the current thru a known resistance across which the input terminals of the detecting tube are connected, in a manner exactly similar to that described by Dr. van der Bijl. The method of measuring the audio frequency output of the detector tube is somewhat different in that we measure the audio frequency voltage acting on the telephone receivers. This is done by throwing the receivers back and forth from the output circuit of the tube to a potentiometer device (slide wire in Figure 1) which is supplied current from the same alternator from which the alternating voltage in the oscillating tube circuit is obtained ensuring the same pitch. The potentiometer voltage is adjusted until equality in sound intensity is obtained. The potentiometer current and setting then give the audio frequency voltage acting on the receivers. This method involves no changes in the electrical characteristics of the plate circuit of the detector tube. Since it likewise permits a continuous variation of the voltage acting on the telephones, it therefore furnishes the data to answer Mr. Ballantine's question as to the accuracy of such an intensity comparison.

We find that audio frequency emfs. can be compared in this way to about one-half of one per cent. when the mean of a number of observations is used.

There is one point in connection with the use of the constant impedance receiver shunt which can possibly introduce a slight error in some measurements. My idea of this device is that when the shunt is varied, the impedance, as measured from the input terminals of the shunt, is constant, but its angle varies. Thus, in one position, when the receivers are unshunted, the impedance can be represented by a vector which makes an angle of about 70° from that which would represent a pure resistance, assuming the values for the receiver impedance and resistance given in the paper. As the shunt is varied, the angle decreases, the modulus remaining constant. In case the shunt is used in the plate circuit of a vacuum tube, it will be in series with a pure resistance. The total circuit impedance will be given by the vector sum of these separate impedances and will, therefore, vary as the shunt is changed.

H. J. van der Bijl: Dr. Miller assumes a large angle for his receiver (82°). Generally receivers have angles considerably smaller that this value. If you calculate the shunt and series resistances of the receiver shunt, assuming that the receiver connected to it has an angle of 40°–50°, you will find that for current attenuations greater than that corresponding to about 6 miles (9.6 km.) of standard cable, the effect of the angle is negligible. This is because for large attenuation the series resistance is large and the effective angle therefore small. In this case we can, therefore, neglect the effect of the angle, which makes the computation of the necessary resistances very simple. If, on the other hand, the angle of the receiver is large, it must be taken into consideration and this has been done in the receiver shunts used in my experiments. In this case the problem is that of an inductive circuit with varying angle. The computation is somewhat tedious and therefore we generally determine the resistances graphically.

I was interested to hear about the experiments made by Dr. Miller at the Bureau of Standards, using this method. I could not quite gather from what he said whether he had made marked improvements on it; I understand that he made measurements which gave an accuracy of one-half of one per cent. I do not exactly know what that means.

I am very glad to hear that this method has been found to allow of such accuracy by Dr. Miller in the comparatively short time. Altho I disclosed my method of measuring detecting efficiency to the Signal Corps about two years ago, it was only about six or eight months ago that I acquainted the Bureau of Standards with the details of it.

HARMONIC OSCILLATIONS IN DIRECTLY EXCITED ANTENNAS USED IN RADIO TELEGRAPHY*

By

Luigi Lombardi

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Introduction

Direct excitation, originated by Hertz, in his classic experiments, and subsequently used by G. Marconi in the first radio telegraphic stations, has meanwhile been practically wholly eliminated in modern stations of any importance, since the advantages afforded by indirect excitation by means of electromagnetically coupled circuits have come to be appreciated. Investigators in radio telegraphy have consequently never made a very careful examination into the oscillations produced in antennas by the more primitive method; and while they did become aware of the complex nature of the emitted waves, they confined themselves in most instances to the observation or calculation of the length of the first harmonic, without determining the amplitude; (rightly considering the harmonic oscillation to be merely an undesired source of loss, inasmuch as the energy radiated thereby was practically in no instance utilized for the transmission of signals, and contributed not towards the efficiency of such transmission, but towards increasing the losses inherent in the entire system).

From a theoretical point of view, moreover, the problem could be considered as having been completely solved by the well-known equation for the propagation of currents in circuits having distributed inductance and capacity, the general integral of which has been given in a sufficiently simple form by K. W. Wagner, in a classic treatise by that author, who also interpreted from a practical point of view, in the same volume,[1] the most important results obtained with reference to aerial and underground lines.

* Received by the Editor, November 12, 1918. Translated from the Italian and edited by the Editor.

[1] "Elektromagnetische Ausgleichsvorgänge." 1908.

Using as a model a conductor or line of considerable inductance and capacity, the same author carried out experiments relative to the transient phenomena, following sudden changes of current and potential, and demonstrated that the actual phenomena were exactly in agreement with the theoretical conclusions arrived at with regard to the same[2]. The effects produced were shown in a number of oscillograms. Wagner's artificial line, however, applied only to low voltage experiments, in which the making and breaking of the circuits occurs without any material sparking. The circuit resistance and the damping being consequently practically constant, the phenomena may be considered to be correctly represented by the theoretical formulas, containing the heat losses and ohmic drops, while radiation need not be considered as of any importance.

On the other hand, the antennas used in radio telegraphy are very different in this respect, inasmuch as the radiated energy in some cases is about as large as the energy dissipated by the ohmic resistances. Among these latter resistances (at least whenever the excitation is accomplished by means of devices utilizing disruptive discharges), the most important one in many cases is that of the spark, which also gives rise to a damping factor, which is rather indefinite and exceedingly variable.

We have recently constructed at the Electrical Engineering Institute of the Royal Polytechnic School an artificial line for high voltages (which line can also be operated as a powerful Hertzian oscillator, or as a radio telegraphic antenna of low radiation resistance) and we have available means for generating radio frequency currents. It was decided to investigate the operation of this apparatus with oscillations directly excited by means of spark discharges of the principal types still used in radio telegraphic stations. Such spark discharges had been recently used by me in other research work concerning excess voltages and the protective systems.[3]

General Experimental Arrangement

For these measurements, the antenna was composed of two large solenoids of copper wire, 3 millimeters (0.12 inch) in diameter and 170 centimeters, (67 inches) long, each of the solenoids having 240 turns 36 centimeters (14.2 inches) in diameter, with an ohmic resistance of 2.7 ohms and an inductance of 0.004 henry.

[2] "Elektrotechnische Zeitschrift," 1911, page 899.
[3] "L'Elettrotecnica," 1918, "Conferenza sperimentale alla A. E. I."

This artificial antenna is shown in Figure 1. A is the 42- or 150-cycle alternator, T the transformer, M the Marconi synchronous rotary gap, a an alternative fixed gap, b coupling coil to the wave meter, and L and C the artificial antenna. The constants are indicated in the diagram.

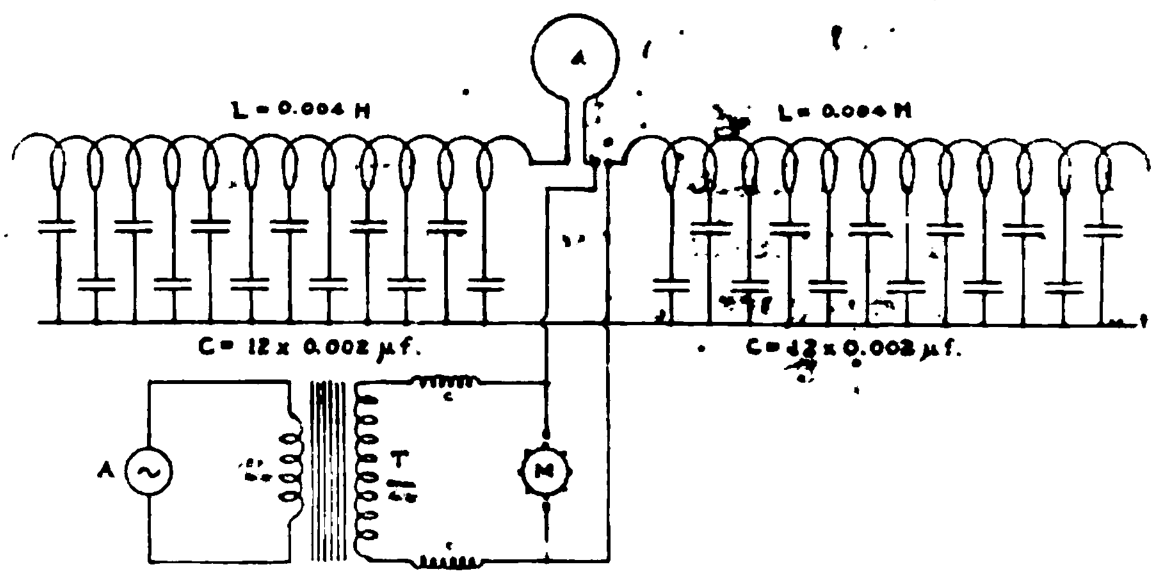

FIGURE 1—Artificial Antenna and Exciting System

The capacity of each section of the antenna was supplied by 12 Moscicki condensers having a total capacity of about 0.026 microfarad, one of the terminals of each condenser bank being grounded while the other was connected to its solenoid at distances of 20 turns from each other. With this arrangement, the fundamental wave length is about 11,000 meters, when the two sections are symmetrically excited at their connecting point; the surge impedance at this wave length proved to be 395 ohms, and the damping constant 330.

The antenna was sometimes excited by means of a large induction coil, with hammer break or turbine interrupter, or else with a Wehnelt interrupter, and with a spark gap having brass or zinc spherical terminals. At other times, excitation was accomplished by means of a 5-kilowatt transformer of commercial type, fed with 42-cycle alternating current from the city supply or from a small 150-cycle alternating current generator with rotating field. Keyed to the shaft of this latter alternator was a synchronous rotary spark gap of the Marconi type, with adjustable fixed terminals, controlled as to exact position by means of a worm gear attachment. In place of the rotary gap, the transformer may be connected to an ordinary spark gap having brass balls, or else to a Boas multi-section spark gap provided with tungsten electrodes, powerfully cooled, and of the type used in radio transmission by Wien's quenched spark system.

For measuring wave length, there was used a Marconi standard wave meter, provided with four fixed capacities and one variable capacity, and with a range of wave lengths from 400 to 6,000 meters. Shorter and longer waves could be measured in the usual way, with added capacity and inductance.

A diagram of the wave meter used is given in Figure 2. L_1 is the inductance, T_1 and T_2 the telephone terminals, H the crystal detector, C the variable condenser, C_1 and C_2 terminals for an external larger condenser, K_1 and K_2 keys, L_2 and L_3 extra inductances which would change the wave length by the same percentage (different in each instrument), R a resistance of 3 ohms, *ab* terminals for decrement measurement, and *cd* terminals for thermocouple. With K_1 to left and K_2 to left (normal settings), L_2 is in circuit. Throwing K_1 to right cuts L_2 out of circuit, while throwing K_2 to right and K_2 to left puts L_3 into the circuit.

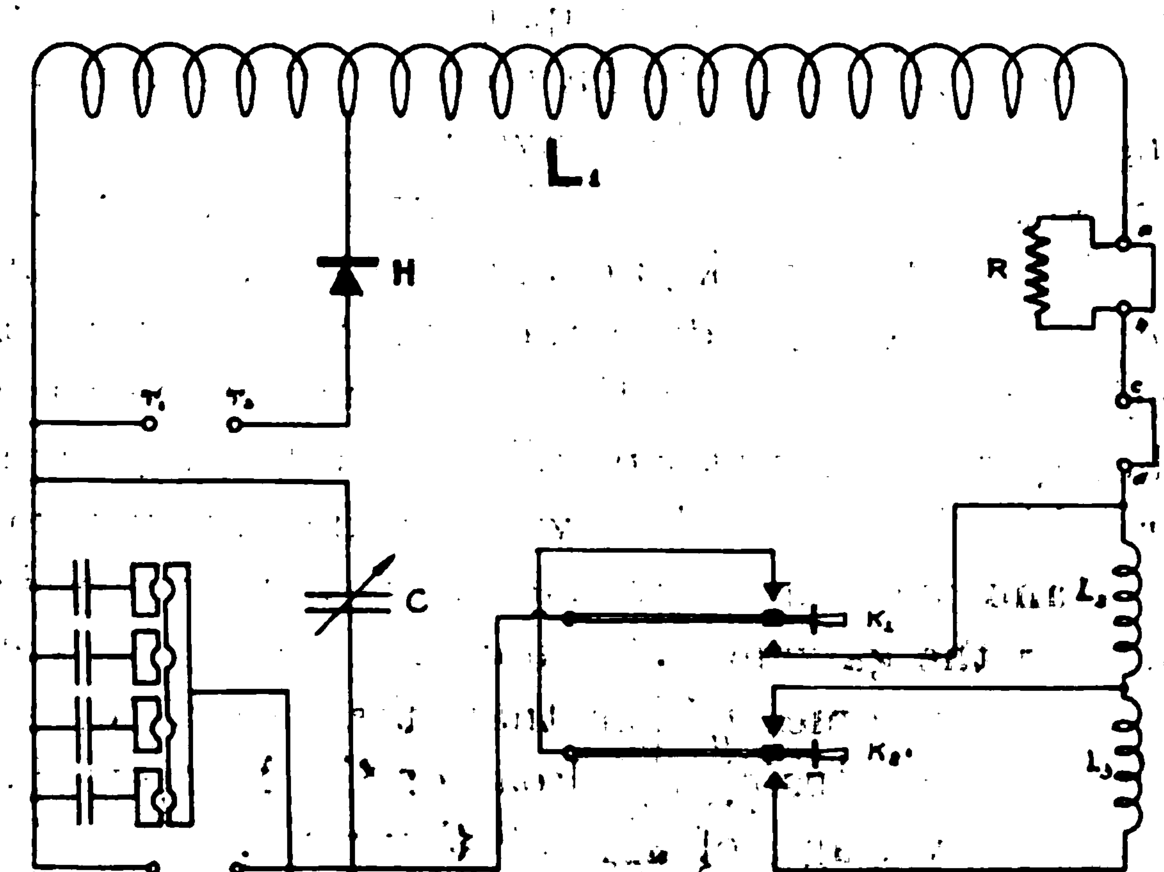

FIGURE 2—Wavemeter Used in Measurements

The wave meter is provided with two supplementary inductances, which could be switched into or cut out of the circuit by means of suitable keys, thus increasing or decreasing the frequency by 3.7 per cent. By means of a thermocouple and a precision galvanometer, two points of the resonance curve, symmetrical with respect to the resonance point, can thus be quickly determined, and the data thus obtained are sufficient for determining the total damping of the combined primary

and secondary. For determining the primary and the secondary decrements separately, a supplementary resistance is included in the wave meter, and furnished with terminal clamps and a short-circuiting bar, by the removal of which bar the total resistance in the circuit is increased by a known amount, while the total damping is increased by an amount dependent on the wave length, and calculable in advance.

Using this method, only a few readings of the wave meter, kept in one position, are necessary to determine the primary current amplitude, containing, however, a factor of proportionality which depends only on the coupling coefficient to the wave meter.

If the position of the wave meter remains unchanged during a series of measurements at different wave lengths (or if, in case the meter has to be moved to a second position, by means of some separate measurements, the relative degrees of sensitiveness in both positions are determined by comparison), it becomes possible to determine quickly the amplitudes of the primary current and of all the harmonic currents in the range of wave lengths covered by the wave meter.

There were connected into the artificial antenna (double solenoid) close to the spark gap, and at a suitable distance from the solenoids (so as to reduce their direct influence), a couple of rectangular turns, similar to the inductances of the wave meter.

By sliding the wave meter parallel to itself along a supporting board or table, I succeeded in varying the coupling coefficient between it and the "antenna" circuit within very wide limits, and in having the galvanometer indicate perceptible deflections even for the harmonics higher than the twentieth, when the antenna was excited using the Boas or the Marconi discharger at a maximum voltage of several thousand. When using a brass spark gap, the measurements on the harmonics became comparatively unreliable; indeed the spark length had to be increased, with the result that the spark became rather unstable; or else the coupling had to be made so close as to cause distortion of the resonance curve. I refrained for the same reason from making a quantitative determination of the harmonics higher than the eighteenth. However, when using the Marconi gap, which operated more regularly, the presence of such harmonics was plainly indicated by the wave meter. Another wave meter having a small inductance, and constructed at our Institute, permitted measurements of shorter waves when the amplitude was reduced to a few per cent. of that of the funda-

mental wave. However, the determination of the new factors thus introduced was unreliable.

According to Wagner's theory, subsequently developed by Petersen[4], and recently re-investigated by Brylinski[5], the discharge of a line of length l, originally charged to the uniform potential E, and accidentally grounded at one end while the other end remains insulated, will be an oscillating phenomenon, equivalent to the superposition of two waves of rectangular form or profile, of a height $\frac{E}{2}$ and a length $4l$, moving in opposite directions with a velocity $\frac{1}{\sqrt{L_1 C_1}}$ in which fraction L_1 and C_1 represent the inductance and the capacity per unit of length. The moving double wave would maintain its initial amplitude for an indefinite time if there were no ohmic resistance in the line, and in the absence of leakage, radiation, and any other factor causing loss of energy. If any of these factors occur, the travelling wave will become attenuated in course of time, and the decrement, calculated for the duration of a cycle, will remain constant if the ratio of the ohmic resistance to the inductance likewise remains constant. An increase in the resistance will therefore result in a more rapid decrease in the amplitude of the travelling wave, and will reduce its total duration, leading up to the new electrical conditions of the system.

Analytically, the rectangular double wave of discharge may be represented by two series of infinitely extended sinusoidal waves of lengths equal to $\frac{4l}{1}, \frac{4l}{3}, \frac{4l}{5}$, and of amplitudes $\frac{2E}{\pi}, \frac{2E}{3\pi}, \frac{2E}{5\pi}$, travelling in opposite directions and with equal velocities. Therefore, if the damping were eliminated, the amplitudes of the successive harmonics (as compared to that of the fundamental wave) would decrease in an inverse arithmetic progression of odd numbers. Consequently no wave corresponding to an even harmonic could be produced, because of the well-known property of functions symmetrical in their successive half-periods. In the special case considered, however, it would not be allowable to assume beforehand that there are no such waves corresponding to even harmonics present inasmuch as the amplitude of the various waves will in-

[4] "Archiv für Elektrotechnik," volume 1, page 233; "Elektrotechnische Zeitschrift," 1913, page 167.

[5] "Revue Générale de l'Electricité," 1918, page 43.

evitably decrease in *geometrical* progression with increasing lapse of time and distance covered, even assuming that the insulation is perfect and the leakage zero. Consequently there will be an alteration in the form of the composite wave in the course of the successive half-periods, and the straight line which gave the outline of the positive and negative maxima of the rectangular wave at the beginning, will be replaced by other (logarithmic) curves. The less pronounced the asymmetry of the half waves (resulting from the damping) proves to be, the less will be the amplitudes of the even harmonics, and the more closely will those of the odd harmonics approach the amplitudes theoretically determined for the rectangular wave.

In Wagner's experiments, conducted with the previously mentioned low voltage line, the decrease of the wave was practically entirely due to the ohmic resistance of the coils, and since the resistance was rather limited $\left(\frac{R}{2L}=12.8\right)$, the decrease in wave amplitude did not amount to more than 15 per cent. after passing over the full length of the line. In the case of the antenna used by me for my research, the decrement was about 25 times as large because of higher ohmic resistance. However, since the natural period of oscillation is 250 times less, the wave should approximate more closely to the rectangular form in this case. Yet, since the antenna had to be excited directly by means of high voltage spark gaps, an additional damping factor was introduced by the spark; and this factor varied with the spark length of the latter, the current thru the gap, and the type of discharger used. For these reasons, the effects produced could not be theoretically predetermined, or at best only qualitatively. This induced me to undertake this modest quantitative research.

Results of Experiments

Of the many experiments made, only a few will be described in this paper, but they will be sufficient to show the general nature of the influence produced by various types of discharger and different operating conditions.

In Table 1, for example, are given two sets of measurements, taken using the Marconi spark gap with 5,000 volts effective, or 7,000 volts maximum, across it. In the first case, the stationary terminals were set to a small spacing of 0.5 mm. (0.02

inch) from the moving electrodes, while in the second case a greater separation of 2 mm. (0.08 inch) was used.

The wave lengths are given in the first column, as determined by the wave meter, without making any corrections for such slight errors as were found because of the irregular distribution of the effective capacity, which would partly explain the discrepancy in the values of the frequency of the successive harmonics as compared to exact multiple values of the fundamental frequency.

In the second column have been entered the numerical ratios of these frequencies (which are inversely proportional to the wave length); and this leaves no room for doubt as regards the absence of odd harmonics and the presence of some even harmonics (and the existence of some other harmonics of an intermediate order), inasmuch as the same results were obtained in many different series of experiments, made under different conditions, and verified by means of various wave meters. The comparison between the actual discharge wave and the typical rectangular wave is simplified, by multiplying the frequency of some of the harmonics by the amplitude of the corresponding current as determined by calculation. In the typical rectangular wave, the resulting product should be constant for the odd harmonics, while for the even harmonics it should reduce to zero.

In the following columns of the table have been entered the logarithmic decrement of the antenna circuit *per complete period*, and the product of the amplitudes of the primary currents (calculated for each of the harmonics in accordance with the formula of Bjerknes) multiplied by the ratio of corresponding frequency to that of the fundamental wave. All the previous data are given for two different settings of the spark gap, namely with large and small sparking distances between electrodes.

The fact that the changes in the decrement are not perfectly regular, is presumably to be attributed partly to unavoidable errors of measurement and partly to the variable resistance of the spark. Moreover, these same factors also influence unfavorably the exact determination of the amplitudes; and therefore the calculated figures show discrepancies between different groups of experiments that are not wholly negligible. Their average is important in this research, since it shows in the most compact form the general character of the phenomenon.

The above remarks likewise apply to the data in Table 2, which represent the results of the measurements taken with the spark gap having spherical electrodes, and with the Boas

spark gap with tungsten electrodes. In both these gaps the spark passes between stationary electrodes, but in consequence of the different composition and arrangement of the terminals, the sparks in the two cases have very different persistencies, which fact is plainly evidenced by the difference between the respective decrements, in turn producing different wave forms.

TABLE I

EXPERIMENTS CONDUCTED WITH THE MARCONI SPARK GAP

		Short Gap		Long Gap	
λ	$\frac{\lambda_1}{\lambda}$	δ	If	δ	If
10,800	1.0	0.300	100	0.440	100
3,530	3.1	0.074	45	0.200	63
2,020	5.3	0.054	36	0.184	61
1,320	8.2	0.054	40	0.172	74
960	11.2	0.056	44	0.160	81
750	14.4	0.056	40	0.144	63
610	17.7	0.058	45	0.120	57
Average Values		0.058	42	0.164	66

TABLE III

EXPERIMENTS CONDUCTED WITH THE BRASS SPARK GAP AND WITH THE BOAS GAP

		Brass Gap		Boas Gap	
λ	$\frac{\lambda_1}{\lambda}$	δ	If	δ	If
10,600	1.0	0.170	100	0.82	100
3,500	3.0	0.166	65	0.26	75
1,920	5.5	0.124	80	0.24	74
1,290	8.2	0.100	95	0.20	109
945	11.2	.....	...	0.18	119
745	14.2	.....	...	0.16	135
612	17.3	.....	...	....	
Average Values		0.130	80	0.21	102

By reason of the increasing irregularity of the phenomenon, at the higher harmonic frequencies the decrements and the amplitudes ascertained with the gaps described are less reliable, and some of them have accordingly been omitted in the Table.

However, a comparison of the results obtained appears to justify the following conclusions:

In radio telegraphic antennas, directly excited by means of an ordinary metallic electrode gap fed from a high voltage source of continuous or alternating electromotive force, the discharge wave has a complex form, which depends on the original potential distribution and on the decrements which arise from the dissipation of energy. Among the sources of loss, the resistance of the spark is of very material importance. The spark resistance depends on the nature and arrangement of the electrodes, by which the ratios of the amplitudes of the upper harmonics to that of fundamental waves are materially modified. The travelling wave has rather the form of Figure 3b than of Figure 3a.

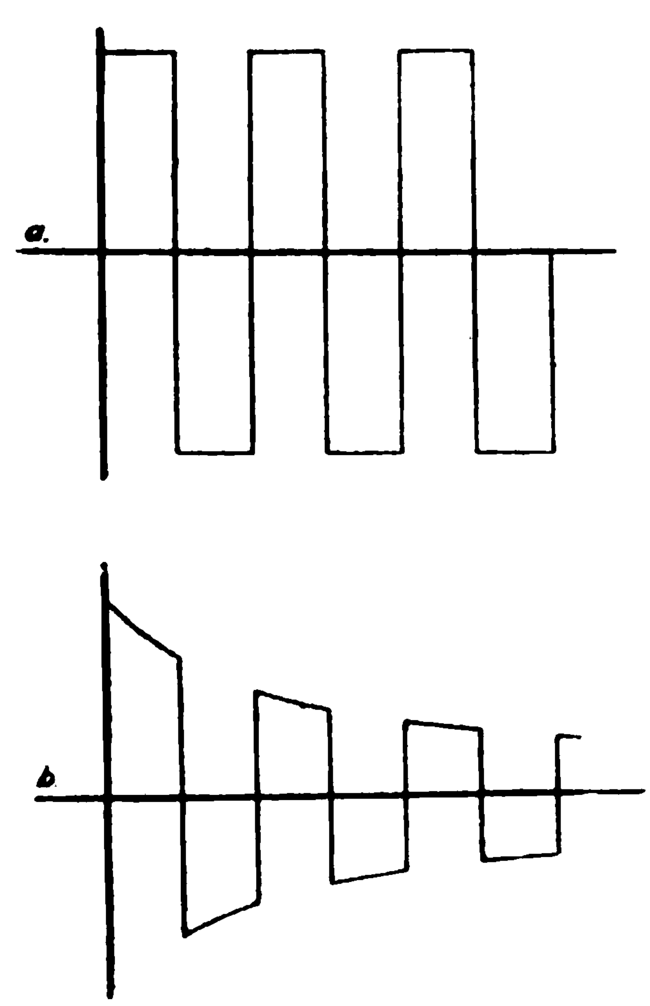

FIGURE 3

The relative amplitudes and decrements, and the frequencies of each of the harmonics can be satisfactorily determined by means of the wave meter which, while it does not afford means for the determination of their relative phases, neverthe-

less allows us to make a useful comparison of the actual harmonics with the corresponding components of the rectangular wave which would theoretically be produced in the case of a uniform original distribution of potential, and with zero damping.

Measurements thus taken on an antenna of low radiating capability made it possible to prove the presence of such harmonics up to, and even beyond the twentieth, and to determine the amplitude and the decrement of each as far as the eighteenth for some types of dischargers. The frequencies found are partly those of the odd harmonics, and partly of the even harmonics (or harmonics of an intermediate order) which latter are due to the dissymmetry of the waves produced during successive semi-periods.

The decrement measured generally diminishes as the frequency increases, but at a rate which is less than proportional to the reciprocal of the latter, and the damping factor therefore apparently increases with an increase in the frequency. The relative amplitude decreases in proportion as the frequency increases, but the product of the two factors which, in the case of the rectangular wave, should remain constant for the odd harmonics, and zero for the even harmonics, actually is slightly variable for the types of gap in which the spark ceases after a few oscillations, while it gradually increases in the other types of gap which produce a more persistent spark. In spark gaps of any given type, the average value of the said product will therefore increase with increase in distance between the electrodes and resulting increase of the damping.

The foregoing experiments are chiefly related to the form of the discharge wave as found at the base of the radio telegraphic antenna, or at the center of the Hertzian oscillator, and where—if the damping is zero—the potential node should be, as well as the common anti-node of all the stationary-wave currents corresponding to the fundamental oscillation and the upper harmonics.

The form of the wave is modified, however, from point to point along the antenna by the relative differences of phase of the harmonics, which directly influence the current and potential variations, as theoretically indicated.

In the case of the antenna herein described, and with the wave meter coupled to one of the ends thereof, the experiments showed with certainty the presence of the odd and even harmonics, as found in the center; and also showed the presence of even

harmonics of a lower order, namely second and fourth, as well as some of higher order, which were missing in the center. However, I have not carried out a quantitative investigation of these latter, the interpretation of the results of which would have been a laborious task, particularly in view of the different coupling coefficients relating to the measurement of each of the harmonic currents in the circuit.

SUMMARY: Using a symmetrically excited artificial antenna consisting of two long coils with a spark gap between them, the author measures the frequencies, decrements, and the relative amplitudes of the fundamental current and of each of the harmonics. It is found that the theoretical rectangular wave, travelling along the antenna, is modified in form, particularly if the spark is fairly persistent. Numerical data are given.

RE-ENFORCED HARMONICS IN HIGH POWER ARC TRANSMITTERS*

By

F. A. Kolster

(Physicist, Bureau of Standards, Washington, District of Columbia)

(A Discussion of a Paper by Luigi Lombardi, on "Harmonic Oscillations in Directly Excited Antennas")

In reading over Professor Lombardi's interesting paper, I am reminded of some experiments which I had occasion to make at the San Francisco transmitting station of the Federal Telegraph Company in the fall of 1914, for the purpose of determining the cause of extraordinary interference produced by this arc station at a wave length much shorter than the fundamental wave transmitted.

High power arc transmitting stations as equipped by the Federal Company consist of a directly excited antenna system; that is to say, the arc is placed directly in the antenna and the length of the wave emitted is determined entirely by the constants of the antenna system. Now it is well known that the antenna circuit is one having several degrees of freedom and when excited directly will therefore emit waves of shorter length than the fundamental, but of much less intensity. In fact, it is not expected that the intensity of these shorter secondary waves would be sufficient to produce any serious interference.

The arc generator produces in any oscillating circuit oscillations which contain to a greater or lesser degree all even and odd harmonics of the fundamental oscillation. If the oscillating circuit which is supplied by the arc generator is one of several degrees of freedom, then it is possible, as will be seen later, to obtain a marked re-enforcement of at least one of the arc harmonics.

Let us consider the theoretical cotangent character reactance curve of the ordinary antenna as shown in Figure 1. Points of zero reactance are obtained at ωf corresponding to the fundamental wave length of the antenna, at $\frac{\omega f}{3}$ at $\frac{\omega f}{5}$ and so on.

* Received by the Editor, April 1, 1919.

When a loading coil is inserted in the antenna, the zero reactance points are obtained at ω_o, corresponding to the fundamental or main transmitted wave, at ω_1, at ω_2, and so on. One would expect, therefore, to obtain re-enforcement of arc harmonics corresponding in frequency to ω_1, ω_2, and so on.

Experimental observations obtained at the South San Francisco stations of the Federal Company confirmed the above predictions as well as could be expected.

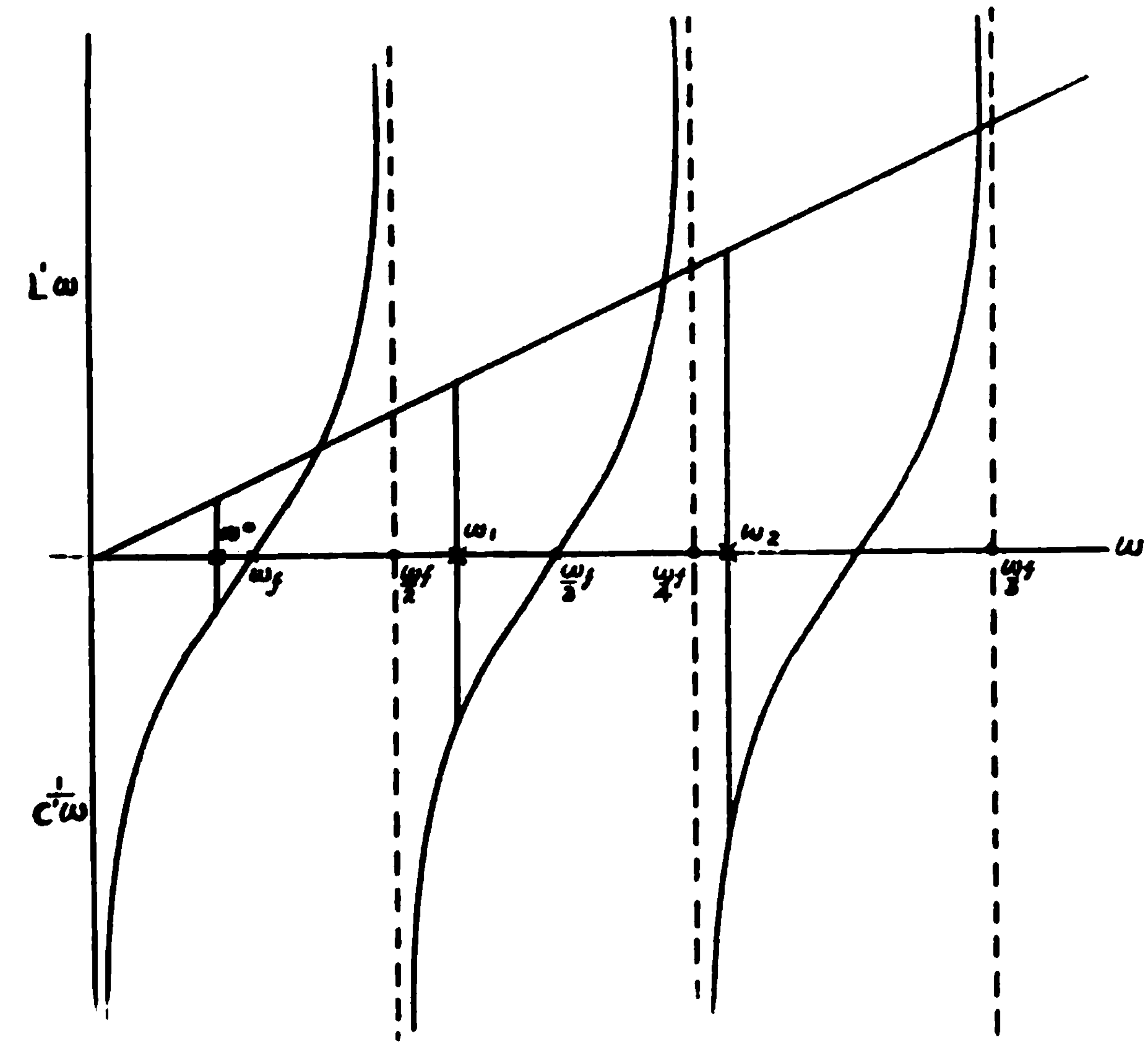

FIGURE 1

From measurements made upon the South San Francisco antenna, the reactance of the antenna was found to be fairly well expressed by the following equation:

Reactance $= 230$ cotangent $(107\,\omega \times 10^{-6})$

This is plotted in Figure 2, together with the reactance of the loading coil inserted in the antenna and it is seen that two points of zero total reactance are obtained, one at ω corresponding to 7,200 meters and another at ω corresponding to about 1,100 meters. The wave lengths corresponding to the odd and even harmonics of the arc are as follows:

7,200 = fundamental	1,440
3,600	1,200
2,400	1.028
1.800	900

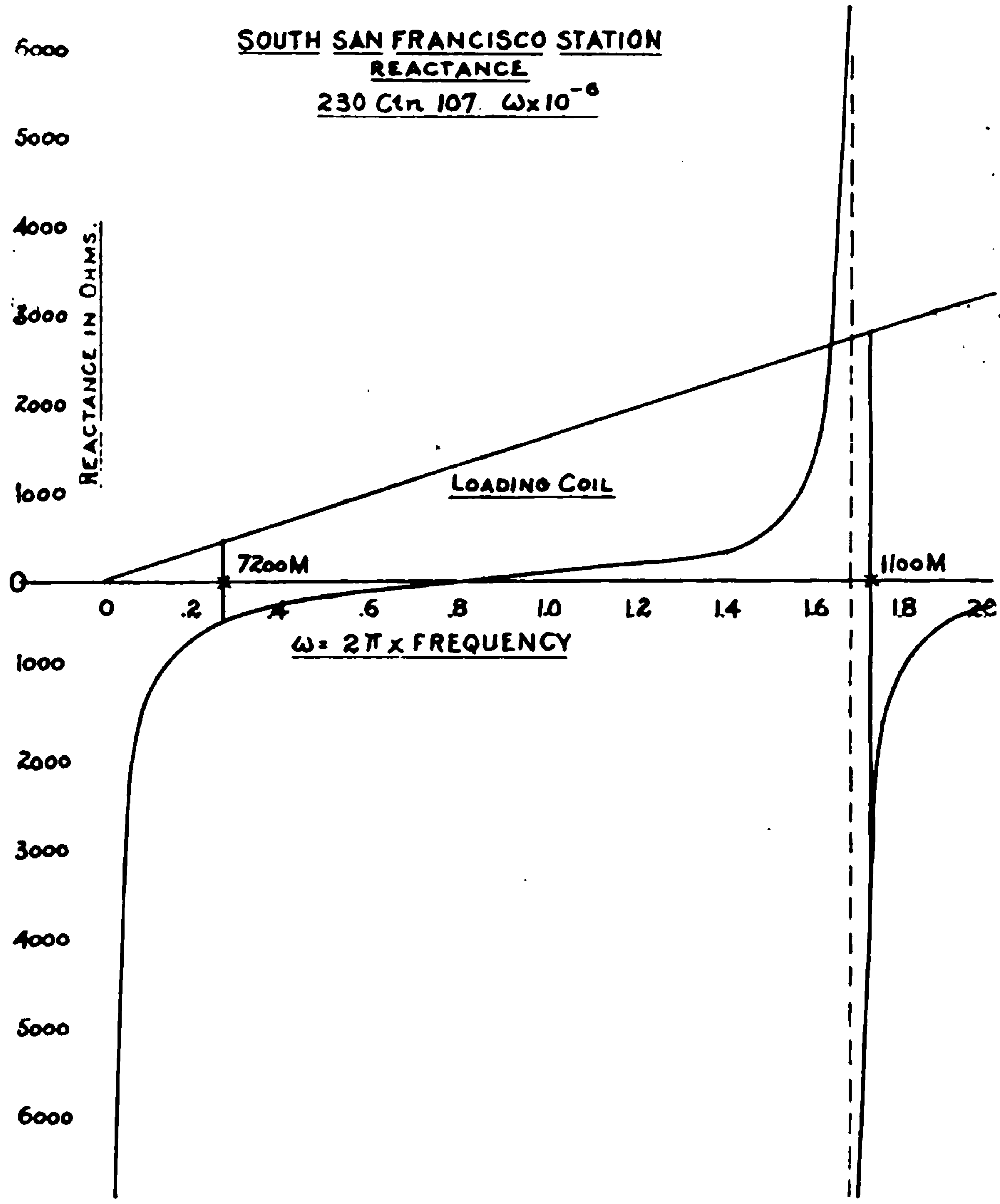

FIGURE 2

The extraordinary interference experienced occurred between 1,000 and 1,200 meters, and tests made at the transmitter showed that considerable energy existed in the antenna in this range, in fact, it was possible to light up three tungsten lamps connected in series with a circuit tuned to approximately 1,100 and coupled to the antenna circuit as shown in Figure 3. Furthermore, a wavemeter containing a hot wire indicating instrument coupled very loosely to the antenna circuit gave full deflection at about 1,100 meters whereas at other wave lengths corresponding to arc harmonics no deflection whatever could be obtained. In other words, due to the characteristics of the antenna system a marked re-enforcement of the 5th or 6th arc harmonic occurred and the energy radiated at this wave length was sufficient to cause serious interference at short wave receiving stations located at a considerable distance away.

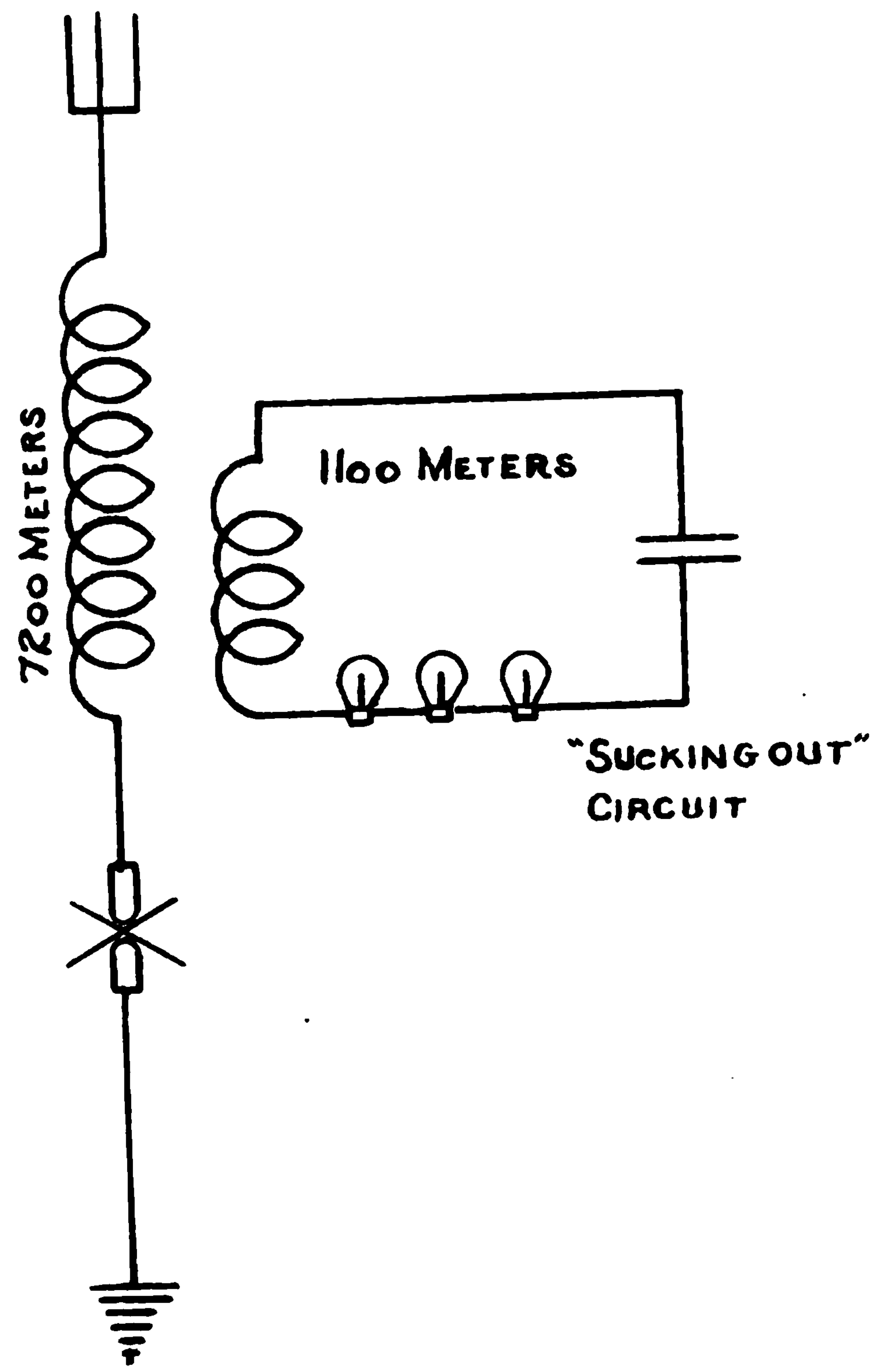

FIGURE 3

It is believed that a "sucking out" circuit, as shown in Figure 3, will materially reduce such interference and as high power arc stations multiply in number it is probable that some means will have to be devised to reduce the disturbing effects of these re-enforced arc harmonics.

SUMMARY: **A Poulsen arc in the antenna will give rise to a series of non-harmonic oscillations of shorter wave length, some of which may cause troublesome interference with nearby short-wave receiving stations. The effect is explained, and a method of reducing such interference by the use of a suitably tuned absorbing circuit coupled to the transmitting antenna is given.**

FURTHER DISCUSSION* ON "ELECTRICAL OSCILLATIONS IN ANTENNAS AND INDUCTION COILS" BY JOHN M. MILLER

By

John H. Morecroft

I was glad to see an article by Dr. Miller on the subject of oscillations in coils and antennas because of my own interest in the subject, and also because of the able manner in which Dr. Miller handles material of this kind. The paper is well worth studying.

I was somewhat startled, however, to find out from the author that I was in error in some of the material presented in my paper in the Proceedings of The Institute of Radio Engineers for December, 1917, especially as I had at the time I wrote my paper thought along similar lines as does Dr. Miller in his treatment of the subject; this is shown by my treatment of the antenna resistance.

As to what the effective inductance and capacity of an antenna are when it is oscillating in its fundamental mode is, it seems to me, a matter of viewpoint. Dr. Miller concedes that my treatment leads to correct predictions of the behavior of the antenna and I concede the same to him; it is a question, therefore, as to which treatment is the more logical.

From the author's deductions we must conclude that at quarter wave length oscillations

$$L_e = \frac{L_o}{2} \tag{1}$$

and

$$C_e = \frac{8}{\pi^2} C_o \tag{2}$$

The value of L really comes from a consideration of the magnetic energy in the antenna keeping the current in the artificial antenna the same as the maximum value it had in the actual antenna, and then selecting the capacity of suitable value to give the artificial antenna the same natural period as the actual antenna. This method of procedure will, as the author states, give an artificial antenna having the same natural frequency, magnetic energy, and electrostatic energy, as the actual antenna, keeping the current in the artificial antenna the same as the maximum current in the actual antenna.

* Received by the Editor, June 26, 1919.

But suppose he had attacked the problem from the viewpoint of electrostatic energy instead of electromagnetic energy, and that he had obtained the constants of his artificial antenna to satisfy these conditions (which are just as fundamental and reasonable as those he did satisfy); same natural frequency, same magnetic energy, same electrostatic energy and the same voltage across the condenser of the artificial as the maximum voltage in the actual antenna. He would then have obtained the relations

$$L_e = \frac{8}{\pi^2} L_o \tag{3}$$

and

$$C_e = \frac{C_o}{2} \tag{4}$$

Now equations (3) and (4) are just as correct as are (1) and (2) and moreover the artificial antenna built with the constants given in (3) and (4) would duplicate the actual antenna just as well as the one built according to the relations given in (1) and (2).

I had these two possibilities in mind when writing in my original article "as the electrostatic energy is a function of the potential curve and the magnetic energy is the same function of the current curve, and both these curves have the same shape, it is logical, *and so on.*" Needless to say, I still consider it logical, and after reading this discussion I am sure Dr. Miller will see my reasons for so thinking.

When applying the theory of uniform lines to coils I think a very large error is made at once, which vitiates very largely any conclusions reached. The L and C of the coil, per centimeter length, are by no means uniform, a necessary condition in the theory of uniform lines; in a long solenoid the L per centimeter near the center of the coil is nearly twice as great as the L per centimeter at the ends, a fact which follows from elementary theory, and one which has been verified in our laboratory by measuring the wave length of a high frequency wave traveling along such a solenoid. The wave length is much shorter in the center of the coil than it is near the ends. What the capacity per centimeter of a solenoid is has never been measured, I think, but it is undoubtedly greater in the center of the coil than near the ends.

The conclusions he reaches from his equation (22) that even at its natural frequency the L of the coil may be regarded as equal to the low frequency value of L is valuable in so far as it enables one better to predict the behavior of the coil, but it

should be kept in mind that
when defined as does the auth
terms of magnetic energy an
at the high frequency, is ver
frequency.

One point on which I diffe
is the question of the reactar
nected in parallel, and excitec
natural frequency of the circ
ance as infinity at this frequ
When the impressed frequenc
frequency there is a high capa
slightly lower than resonant f
reaction, but at the resonant f
cuit is zero. The resistance
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but for any value of coil resi
bination is zero at resonant fr

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